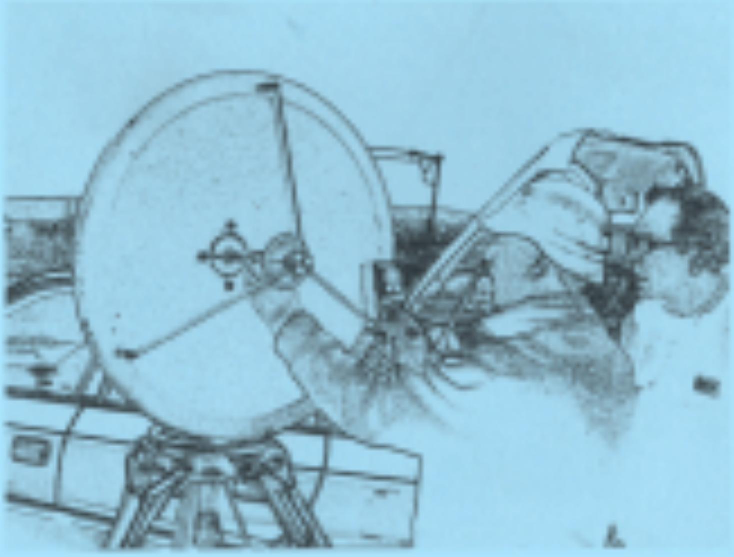
UK Microwave Group



BACKSCATTER

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UK MICROWAVE GROUP

TECHNICAL COMPENDIUM

A selection of technical articles from
Scatterpoint and the Microwave Newsletter
1999-2006

1st Edition: October 2008

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www.microwavers.org

Printed by Mensa Printers of Sheffield

Foreword

It’s not often you find a book with nothing new in it that is actually worth reading.

Well this is one, but the content is not “new” only if you’ve been an avid reader of Scatterpoint and the RSGB Microwave Newsletter over the last 10 years, have a photographic memory, and have lots of bookshelf space. Anyway, where else can you get a handy up-to-date reference that’s all in one slim volume and can sit on your workshop shelf ready for when you refer to it?

When “Scatterpoint” Editor, Peter Day G3PHO asked me if I would take on the mammoth task of putting this together, one of my fellow microwavers used a sentence with the words “chalice” and “poisoned” in it at the same time. Well far from that, it has been an education for me, someone relatively new to Amateur Microwaves, to read and think about the vast amount of knowledge it contains.

You will find contributions from the likes of Dick Knadle K2RIW, whose excellent and thought-provoking “ramblings” on antenna theory and all things microwave kept me amused and fascinated during the preparation process, alongside genuine “nuts and bolts” papers from practical guys like David Wrigley, G8GXK and Bernie Wright, G4HJW who tell us how to build stations from parts you’d pick up at a radio surplus sale for a few pounds.

It’s not just transmitting and receiving equipment though; there are articles on repairing (what used to be very expensive) test equipment alongside great bits of test gear you can build yourself.

I hope you enjoy this and it encourages you to get more out of the fascinating topic of Amateur Microwaves.

John Worsnop 2008

CONTENTS

Frequency sources **4**
  Crystal Oscillators 5
  Synthesisers, PLOs and DDSs 20
  Frequency markers and references 39
  General 50

Frequency Multipliers **58**

Antennas **67**
  Yagis and helixes 74
  Dishes and feeds 87
  Other Antennas 136

Low Noise Amplifiers **147**

Power Amplifiers **169**

Accessories **182**

Transverters and converters **214**

IF rigs **231**

Power Supplies **252**

Keyers and control **263**

Propagation topics **274**

Measurement **298**

Noise **318**

Surplus **335**
  Building blocks 336
  Test equipment 365

Mechanical **392**

Miscellany **398**

Chapter 1

Frequency Sources

Crystal Oscillators

Microwave Oscillator Stability

What to consider in achieving “Best Performance”

John Hazell, G8ACE

This article discusses some of the considerations governing the stability of a quartz oscillator source.

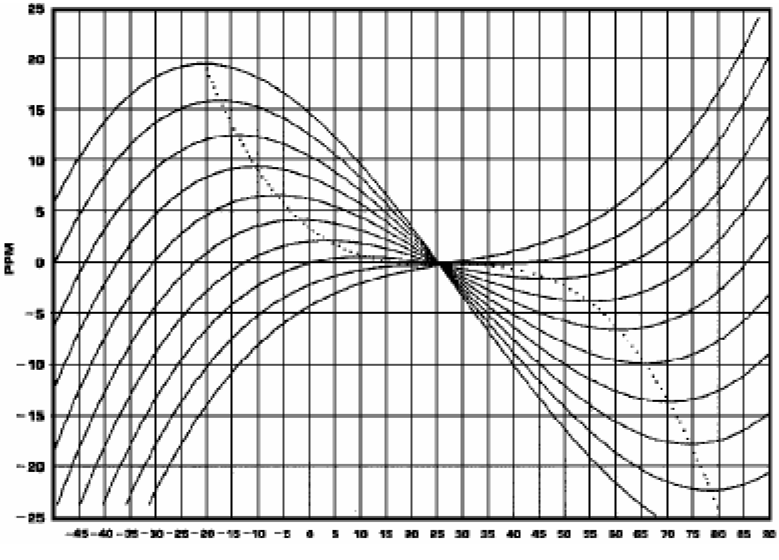
Temperature dependence

Almost without exception, crystals are marked with their operating frequency. However, **operating temperature** is vitally important for stability and, for most crystals, this information is either missing or hidden within a manufacturer’s code. The family of curves in the graph below show the turnover points for AT cut Crystals. The angle of cut governs the turnover or best operating temperature and this angle will typically lie between 34° 58' and 35° 13'.

The ideal condition is to operate the crystal on the flat part of its characteristic. Typically a crystal cut between 35° 0' and 35° 2' would be used at ambient temperature.

Over the wider temperature range of O°C to 60°C it might exhibit a stability of 5ppm or 100kHz variation at 10GHz. The need to control the temperature becomes obvious and to do this the crystal must be held above ambient in a Crystal Oven or Murata-style heater clip. Superior results can be obtained by using a crystal whose angle of cut results in a turnover point matching the heater. Finally, the oscillator circuitry may contain components whose characteristics vary with temperature and hence the frequency will still vary. One solution here is to also temperature stabilise this circuitry.

An ovened oscillator system, set to



1°C of the crystal turnover temperature, would generally provide a high degree of stability. An ovened homebrew 10MHz reference set to 1° of turnover returns better than 1 part in 10**8**/°C stability after six months ageing, or 100Hz/°C at 10GHz.

Ageing rate

A new crystal will typically age at 0.5ppm/month (5kHz shift at 10GHz) with the figure declining substantially after the first few months of continuous operation. Cleanliness in processing, the type of mount and the seal type all affect ageing. The norm is for the frequency to increase in glass enclosures and decrease in metal.

Re-stabilisation

Once a crystal oscillator has been switched off it will take some time to return to a given ageing rate.

Frequency retracing

Again, once the oscillator is switched off, say at the end of an event, and repowered some weeks later, there will be a frequency offset, even after warm up. This may help to explain frequency calibration errors between events.

Considerations

Microwave operators will know that the two most important factors in finding that distant signal are dish alignment and accurate frequency readout. The advantages of oven control on the oscillator are well worth while but, for full advantage, try doing some or all the following:

- In the case of a home constructed oven, measure crystal frequency against thermostat voltage in determining the turnover point of the crystal. Use this voltage as the set point for the oven comparator. Make sure the crystal turnover temperature is above the oven self-heating value so the oven is in control for all ambient operating temperatures.
- Age the oscillator for as long as possible and this generally means months, at least, to see stabilisation of the frequency.
- Avoid switching the unit off. Perhaps diode steering in the power connection arrangements is worth considering so that the oscillator can be moved from mains to battery power without breaking the supply.
- Adequately insulate the crystal oven. When large ambient temperature changes are encountered, an additional outer oven may be required.
- Short term stability ... some crystals exhibit large frequency fluctuations. One cause of this appears to be stress in the connections to the quartz. Heat treatment has been shown to improve this problem.
- Power supply voltage/frequency dependence, shock, vibration and humidity are all other factors to be considered.

Crystal Oscillators

Some ongoing discussion from the UK Microwave Reflector

This article is not the usual formal technical article, but an excellent example of how an internet reflector should be used ...useful and informative discussion between three or four experienced microwavers, posted for all to read. We wish there was more of this kind of discussion on the various reflectors instead of the "back biting" that we sometimes see! So our thanks go to the authors of the posts you are about to read.

Kevin Murphy, ZL1UJG, posted the following on the UK Microwave internet reflector. It started off a most interest discussion as emails went back and forth over several days .....

I am working on some crystal oscillators for beacons. I have a G4DDK004 oscillator and also some Minikits EME65 PCB's, which both use the 2 transistor Butler crystal oscillator. I have looked at the DC across the V (base-emitter) on the 2 transistors (using a 10k resistor to isolate RF from the digital voltmeter and they both are below 0.5 Volts. (below 0.3 Volts in some cases). The voltage (b-e) on the transistors should be 0.6 to 0.7V so that constant source and load terminations are seen by the crystal. I remedied this by putting two 1N4148s across the tuned circuit, which gave 0.58 to 0.7V across each transistor base emitter, with minimal drop in harmonic output.

I added a choke across the crystal to cancel out the stray capacitance of the crystal holder (5 to 7 pF). I additionally raised the impedance of the 1st transistor, by putting a small resistor in series with the base (~ 30 ohms SMD) Measurements indicate that this raises the RF level and should increase the carrier noise, although it will degrade crystal operating Q. I used a resistive 20dB probe to make some comparative level measurements. Although the oscillators should run off a low noise regulator, these ones are run off a bench PSU. The complete oscillator will be run off a heavily filtered 7808 regulator, with the 2 transistor oscillator being run off an active finesse filter. (see www.wenzel.com). I measured the wideband noise and this measures around -160 dB below the carrier (referred back to the crystal oscillator frequency). (The instrument appears to have some margin left. (I can see the roll off of the wideband noise, due to a pipe cap filter used in a 13 cm LO output.)

I looked at the G8ACE website and the wideband noise floor of his oscillator, referred back to the crystal frequency, appears to be of the same order. Then I came across this link: **http://www.nitehawk.com/sm5bsz/linuxdsp/hware/lotest.htm** I understand what is being said and tried a 1uH in series with the emitter resistor of the 1st transistor of the Butler 2 transistor oscillator and, hello! ....I measured no difference!

The comments indicate that the KD6OZH oscillator is really good (-178dBc for wideband noise) (-172dBc in the original article).

Have I missed something? Should I try a FET as the 1st device, since it apparently has less noise? Is the noise generated by the BFS17 (which I have used for the 2 oscillator transistors) that much higher? Why is the G8ACE oscillator apparently not as good as the KD6OZH oscillator? Instrument limitations? Is there another noise source in the oscillators? I know that PSU noise from 78## and 78L## regulators is an issue. I notice that these 2 transistor Butler oscillators, have an extraordinary tuning range, pulling the output frequency +/- many 10s of kHz on the 13cm band, before the crystal oscillator stops. Is this normal? I expect the FET/bipolar version from G8ACE, would have less tuning range as the overall gain in the oscillator circuit is less but is the sensitivity to tuning similar? I know there are probably better oscillators, but I haven't seen any schematics/circuits.

It has been an interesting exercise taking measurements and learning a little more about oscillators.

Comments and discussion please!

Following Kevin's post, the following replies were received…..

From SAM JEWELL, G4DDK, <jewell@btinternet.com>

KD6OZH states in his QEX article that it is important to achieve and retain low noise figure and high linearity in the second stage of the oscillator to obtain the very low noise performance he claims. The use of a cascode second stage can sometimes make this difficult to achieve because the available supply volts is shared between the two devices in series, which might lead to less than optimum performance. My experience is that you need to use a transformer coupled output or a tuned output although the use of low voltage silicon/germanium devices in the cascode might also work.

From John Hazell, G8ACE: The importance of Crystal Temperature Control

The debate on oscillator stability has to date been centred around the Q of the crystal and other parameters of the RF circuit used. Crystals are unstable with temperature and this stability aspect needs addressing just as carefully as the RF electronics. The following comments relate to AT cut overtone crystals as commonly used in microwave equipments. The cutting angle of the quartz controls the knee point. That is the operating point at which the frequency shift is least with temperature change. Common cut angles give knee points of 25C, 40C and 60C although since the angle is so critical some variation from the expected point will be found. 25C is satisfactory for low frequency equipment with little internal heating and has the advantage that the knee slope angle is shallow with temperature changes. 40C is useful for clip heaters but if the equipment temperature rises above the clip temperature then the stability is lost. 60C ensures that for the majority of the time the crystal is above the influence of ambient temperature changes within the equipment. In the G8ACE MKII OCXO part of the setup procedure is to explore for the knee point during the alignment. This is for two reasons: firstly, suppliers do not necessarily supply what is required and prefer their own spec of quoting stability in ppm over a given temperature range. Most cut angles will easily fit into this spec.

It is most important when ordering a crystal to quote for OCXO use and the required turnover point. Secondly, as mentioned above, the precision required for the cutting angle leads to some variation anyway in the actual knee point and only by individual oven adjustment will this point be found with certainty. A guide to the enormity of the frequency change with temperature is that a commonly used 106.5 MHz/60C crystal will move some 800Hz during the heating phase, starting from 15C. Multiplied by 96 for 10GHz this represents some 75kHz of system calibration shift. To confirm this point consider a commercial 5 or 10MHz OCXO where large shifts in frequency occur during warm up but then is quite stable. Once the knee point has been determined, the stability can be good but some crystals do move in frequency very slowly. This is assumed to be part of the ageing process. However, it has been observed that small adjustments to the oven temperature around the knee point affect the amount of frequency creep. For a crystal operated for say 8 hours daily the crystal will retrace overnight and each subsequent day will repeat the process. Powering the OCXO 24 hours a day the frequency creep will slow down but there will be variations between samples. It has been noted that the creep amount is dependent on exactly where the oven temperature is set about the knee point. Nothing has been found in documents relating to this unless the writer has misunderstood that published.

Eric, F1GHB has spent some time also looking into the finer aspects of stability and has kindly sent his results which are shown in page two of his pdf document. For the advanced setup Process, access to either GPs or Rubidium or similar source is necessary. Once the creep rate of the OCXO has been determined the OCXO temperature is adjusted in increments to +/- 0.5v either side of the knee point and the change in the frequency creep noted until the flattest time/frequency response is achieved. In the absence of a high stability source for advanced setup a beacon could be used in the 3cm instance but it must be chosen with care. GB3MHX, whilst free running, is quite stable. GB3SCX is also very stable with Hz errors quoted by Andy G4JNT from time to time. Other beacons driven by Adret sources could also be suitable. This document plus the additional information from Eric F1GHB can be found at these sites: **www.microwaves.dsl.pipex.com/mk2/advanced.pdf** and **www.microwaves.mcmail.com/mk2/advanced.pdf**

From Chris Bartram, GW4DGU

I agree completely with John about finding the inflection/turnover/knee point being the key to good OCXO performance. Trying to significantly temperature compensate oscillators with compensating capacitors is a bit of a nightmare - I know, I've been there! For amateur usage, over a limited temperature range, it's probably possible but it's a slow process, particularly as you have to wait for the oscillator to reach thermal equilibrium every time you change a capacitor. One of the problems with trying to temperature compensate a crystal oscillator is that the frequency/temperature curve isn't linear. Also, you really need some form of controllable thermal environment. Very many years ago, Dave Tong (of Datong fame) wrote a very nice article about compensating VFOs using that technique in 'Wireless World' which very succinctly details how to do it. That can be applied to crystal oscillators. Julian, 'YGF, described a circuit using a thermistor and varactor in Radcom about 15 years ago. There's also a good book on the subject by a guy called Marvin Frerking, if anyone is seriously interested.

For low-power portable operation, I'd probably find a good (0.5ppm) TCXO at a rally (or even buy a new one!!) and lock the DB6NT oscillator to it using one of the techniques that Andy, 'JNT, has recently described. That will give at least an order of magnitude better frequency stability than a simple crystal, and I'd be prepared to lay good odds that it could be done with a current drain of <25mA at 12V. Kevin's measurements of his oscillator are very interesting. He's seeing better performance than I would have expected from my simulations, but, from memory, he's breaking the loop differently, and we may well have used different component values. Mine were lifted from the RSGB Microwave Manual. My simulations of three years ago are now stored on a deep backup CD somewhere.

The important points to note are:

a. He's clearly got a good crystal! But then Rakon (a NZ company) is a VERY good crystal supplier, a world leader, in fact. I've specified them for work projects where excellent crystal performance was required.

b. He shows the reduction in loaded Q with drive-level very clearly.

c. >15dB open-loop gain is grossly excessive and will cause the oscillator maintaining amplifier, operating with the loop closed, to be cut-off most of the time, resulting in very severe degradation of the loaded Q.

Typically, I'd design a crystal oscillator for about +6dB open-loop gain. That would ensure reliable starting. I'd also use a separate limiter, such as a pair of 'crossed' schottky diodes, so that the amplifier is forced to operate in class-A. This could also be used to control the crystal dissipation. There are other ways of making a suitable limiter, such as using a 'long-tailed-pair' (I did work on this many years ago using a Plessey (remember them?) IC limiter, and the idea found its way back into the device data sheet!) but 'crossed' Schottkys are a good solution and much simpler to implement. There are a few points to note about the use of formal limiters in an oscillator design. Firstly, because the amplifier is operating in class-A, large levels of harmonics are not available. I like balanced schottky diode frequency multipliers, as they add almost no additional phase noise. Alternatively resistive mode FET or BJT multipliers can be good but beware of highly efficient BJT multiplier, as they are probably at least partially employing some form of parametric effect, and can act as highly efficient phase modulators, transferring supply-line noise to the carrier. Secondly, it's quite a good idea to take the oscillator output across the limiter, as the amplitude domain noise will be 'squashed'. Don't try to take the output from in series with the crystal - you'll just reduce the loaded Q. There's no free lunch there! A third point to note is that the use of AGC instead of a limiter to keep the amplifier in class-A can easily lead to the generation of phase noise due to the amplitude-to-phase conversion in the gain control device. I hope that's useful to someone.

For what it’s worth, it doesn't come as a surprise that changing a crystal in an oscillator to one on a frequency not so different from the previous one results in the crystal coming-up on frequency! If the oscillator had been set-up to put the frequency of the first crystal on its nominal frequency and the loop phase response of the maintaining amplifier was essentially unchanged, then I'd expect the second crystal to also come-up on frequency, give or take the crystal calibration. There's no reason why the Driscoll oscillator can't be turned into a VCXO. It's entirely possible to tune it, and in an more controlled way an the current alternative. The balloon-board 'Butler' works quite well as a VCXO in a rather uncontrolled manner because the loaded Q of the resonator is relatively low. But, it has disadvantages. In particular, both ends of the varactor diode float above RF ground, and it's quite difficult to get good loaded Q, which means that you won't get the best phase noise, and short/medium term frequency stability. Why do I call the emitter-coupled oscillator, known to the amateur microwave community as a 'Butler' oscillator, the 'balloon board' oscillator? The ancient history is that in the early days of narrowband microwave operation the blessed Mike Walters, G3JVL, was able to procure a significant numbers of a meteorological balloon transmitter which was designed to send data as FSK at about 400MHz. This was an empirical design produced before even most professionals had access time but it was designed to accommodate relatively wide shift direct FSK at UHF, not as a high quality source. It was never designed as an oscillator intended for multiplication to microwave frequencies! Nonetheless, as a convenient source of 384MHz it was used in most of the early 10GHz transverters as a driver for a step-recovery diode multiplier, and formed the basis of a number of subsequent 'designs'. I know, as the result of simulation and analysis, that the 'balloon board' oscillator isn't terribly good. That was clear even with the relatively crude software tools we had 20+years ago.

I've been using both FET and BJT Driscoll-type oscillators for the highest performance applications since the early 1980s, and I've yet to find a better circuit topology for a VHF overtone crystal oscillator. My transverter designs of 20+ years ago for muTek used a single JFET Butler oscillator. (I think DB6NT still uses something similar.) This was acceptable at VHF, but I wouldn't use it at higher frequencies, as it's quite difficult to control crystal dissipation, so it drifts and the loaded Q isn't **that** good. The phase shifts, and thus frequency changes, due to temperature dependent changes in maintaining amplifier component values make it unsuitable for applications which have to meet commercial approvals even at VHF. In other work applications I've used the impedance inverting overtone oscillator with good results and a single transistor version of the Driscoll.

So you think you know all about crystals?

Part 1

From the WA1MBA Microwave Reflector

The following discussion appeared on the WA1MBA Internet Microwave Reflector. We are very grateful to Bob, K3VOT, for allowing us to publish his email information in the form of this article. Bob really knows what he is talking about! While some of his comments are more relevant to American microwavers than to us in the UK, we have not omitted them as they still make interesting reading.

From: K3VOT [Bob@aurand.com]
My qualifications? 25+ years working in the crystal oscillator and microwave oscillator business, including custom microwave bricks, high stability OCXO's and TCXO's, synthesizers, etc. I currently work as engineering manager for a large international corporation in their crystal oscillator division. In a typical year we buy, oh .....400,000 to 500,000 crystals.

Crystal Suppliers:
First of all, I sometimes use International Crystal for home projects, consulting jobs and occasionally for projects with my current employer. I've had excellent success with them, their major shortcoming being lack of capacity—they just can't deal with the kind of large orders we place. Their chief crystal engineer and a good contact there is Darrell Brehm, WA3OPY. Darrell has been around, having worked at McCoy and Reeves-Hoffman before ICM. Some (all?) of the other crystal makers mentioned on this reflector are small outfits and I have no experience with them except Colorado Crystal. Colorado is probably in the top tier of crystal makers, worldwide, and they have an excellent hi-precision process. Contact is Tom Schulyer, KD0JP I believe. In the last 5 years or so, much consolidation and acquisition has occurred in the US crystal industry. Colorado has been one of the survivors and, as a result, has more work than they need or want. Lead times are very long ... for work, we are regularly quoted 30+ weeks.

A good resource for finding crystal and oscillator vendors is the IEEE Frequency Control Society. They have hotlinks on their web site to most of the major suppliers worldwide. Lots of good technical papers here too. I recommend John Vig's tutorial on crystals and oscillators as required reading for all experimenters. The IEEE web page that links to crystal and oscillator related information is:
http://www.ieee-uffc.org/index.asp?page=freqcontrol/freqmain.html&Part=5#top
Crash around in there and you will find lots of good stuff!

Crystal accuracy and behaviour:
Firstly, virtually no one makes solder seal crystals anymore. It's a dirty process, leaving splatters of solder and flux inside the can. Any contamination such as this will move around, eventually landing on the electrodes, changing the frequency. In the biz we call this the quartz to crud ratio. For various reasons the crud can subsequently leave the electrode area only to return later, changing the frequency every time. Additionally, solder sealing is intrinsically an "atmospheric" sealing operation. So, what ever contaminate (or moisture) was inside the building at the time gets sealed inside the can to do

it's damage over time. Modern crystals are invariably either resistance welded or cold welded. Resistance welding is far more common for several reasons: commercial welding equipment can be purchased -- cold welders are usually custom built. R-weld is faster and therefore more suitable for production environments. R-welders are easier to operate and maintain. Among oscillator engineers, it's generally believed that cold welding results in better performance, but some crystal engineers swear resistance welding can achieve the same performance. The jury is still out on that one, as far as I'm concerned. All UM-1's are resistance welded. TO-3 and other 2 leaded crystals can be either R weld or cold weld. Look at the base ... if it's one large piece of glass with the crystal leads poking through, it's cold weld. If the base is mostly metal with matched glass eyelets for the leads, it's R weld. No crystal maker can get the calibration right unless you, the user, supply the right information.

For fixed frequency oscillators (not voltage controlled) this is largely the load capacitance, sometimes called "finish point". "Series" is a legal load corresponding to infinite load cap. Series crystals vs. load crystals are all made the same way. The difference is made up in the finish plating, what someone on this reflector referred to as the final electro deposition of electrode material and is strictly a matter of where on the reactance curve to put the frequency. All crystal makers need some kind of tolerance on load cap (if used) and frequency tolerance. +/- 0.5 pF is typical for load and +/- 10 ppm for frequency. It's up to you to adjust your circuit to frequency (try **THAT** on 400,000 oscillators a year!). Remember, at 1296 MHz, 10 ppm is 12.96 KHz! If the crystal is used in a oven, the operating temperature **must** be specified and given a tolerance. +/- 5 degrees Centigrade is typical. Usually, with OCXO's, the OSCILLATOR maker adjusts the temperature of each oven to match the turnover temperature of the crystal and each crystal in a batch will have slightly different turnover within the tolerance. Don't expect a well-centred process with a Gaussian distribution unless you are buying a VERY LARGE number of crystals. In practice, this means, when you are replacing a crystal in an oven, say to change frequency of a brick, and the documentation says maybe 75 degrees, that's the NOMINAL design temperature for the crystal. Maybe the old crystal came in at 73.4 and the brick manufacturer set the temp on that individual oven there. Now you buy a crystal from say, International, and give them the 75 degree spec but they end up at 78 degrees. When this crystal is installed in the 73.4 degree oven, it will run high and on a slope. If the oven temperature creeps up over time, due to ageing of the thermistor or other components, the frequency will drop towards the turn and this change in frequency will be indistinguishable from ageing.

By the way, this oven setting information applies to AT cut crystals, only. SCs behave differently but are unlikely to show up in ham equipment. NEVER expect a crystal calibrated at room temperature to be on frequency when used in an OCXO. Expect a 20 to 30 ppm drop in frequency (for the same load) when the crystal gets to oven temperature.

Crystal Oscillators circuits:

No oscillator circuit is perfect but by far my favourite is the Butler. It can easily work from low MHz frequencies to VHF. It's easy to set-up, accommodates both fundamental and overtone crystals and can deliver very good phase noise. We run thousands of these using 155MHz fundamental crystals each week ... no problems.

Here's how to do it:
Select a high Ft transistor ... my favourite is the NE856 from NEC and available from Digikey. This transistor has low junction capacitances, low Rbb and good 1/f noise. In fact, I use this transistor in low level stages everywhere from audio to microwave. If you are running from a +12 volt or thereabouts supply, bias the base with 1K ohm to ground and 3.3 K ohm to +V. Chose the emitter resistor to be 1K at low frequencies and maybe as low as 330 ohm at 150MHz. Considerable variation in bias is possible. Bypass the base with a 0.1 uF chip cap close to the base lead. For the tank, select the inductor value to be between 50 and 150 ohms reactance at the oscillation frequency. Wind the coil on a toroid core. Micro Metals T-25-6 or T-30-6 are good choices at 100MHz. Use a larger core at lower frequencies.
Calculate the capacitance to resonate with the chosen inductor. Split the cap in a 1:3 to 1:5 ratio. Again considerable variation is possible. You might need to go back and adjust the inductor value slightly to accommodate standard value caps. Use C0G chip caps at VHF. At lower frequencies where C0G are not practical you will need to go to a higher dielectric constant like X7R for at least one of the caps. Put the larger value cap (smaller reactance) to ground and the other cap to the collector. Initially, take output from a one turn link wound on the toroid.

To set up, connect a jumper or small value resistor from the emitter to the junction of the two tank caps to make the circuit free-run. If necessary, adjust the tank to set the frequency roughly to value. Remove the jumper or resistor and connect the crystal. Observe the output frequency and level and squeeze or stretch the toroid windings if necessary to peak the output level.

Now adjust the output coupling by adding one turn at a time to the previous one turn link, each time squeezing or stretching the toroid winding if needed. While adding turns you will find the output power to increase, level off and then decrease. Proper coupling is fewest number of turns to reach the peak output power. At VHF, one turn may be sufficient. If you must adjust the oscillator frequency, put a setting capacitor in series with the crystal.

If you need a VCXO, put a varactor diode in series with the setting capacitor and the crystal. In the case of a VCXO, the varactor anode goes to the tapped capacitors. A 10K resistor from this point to ground establishes the anode voltage at zero. Another 10K resistor from the cathode provides a tuning voltage input port. Resistor values are not critical ... just don't go too low or too high. The Zetex ZC8xx series varactors, available from Digikey, work pretty well. If pullability is too low you can try some coil in series with the crystal and varactor. Again, wind it on a small toroid and you can have almost infinite adjustability.

The general idea is to make the coil + varactor + setting cap combination approximately series resonant at the oscillation frequency. Pull range is also very dependent on crystal parameters and will ultimately limit the range. If reactances in the series leg get too large loop gain will drop below 1 and oscillation will cease. Make this circuit small, shove it into a piece of 1" diameter copper water pipe, wind the outside of the pipe with insulated resistance wire and you have the basics of an OCXO. I hope this makes sense. A two transistor Butler circuit is possible and has some advantages at low frequencies (like 10MHz). That will have to wait for my web page and real schematics!

So you think you know all about crystals?

Part 2

The reply from Chris Bartram, GW4DGU

While I largely agree with the first 60% of the notes, I'd very strongly suggest that the section about crystal oscillator circuits be renamed 'Here's how NOT to do it', at least for microwave applications!

I've designed a large number of crystal oscillators in the course of my job over the years, many of which my clients have put into large scale production. As part of the design process, I've looked analytically at most of the common designs of vhf overtone oscillator, and come to some conclusions based on hard-headed analysis.

One of the most useful indicators of crystal oscillator performance is the loaded Q of the crystal. The more energy retained in each cycle by the crystal (the greater the loaded Q), the smaller the effect of the maintaining amplifier, and the greater the stability and the smaller the level of the phase noise. Of course it's not quite as simple as that, and the hf noise figure and 1/f noise performance of the maintaining amplifier as well as the power levels and limiting strategies within the oscillator all play their part. To a first approximation, these are all related by simple linear maths. For those who want to go further there are books: Randy Rhea's 'Oscillator Design and Computer Analysis' is a good accessible start. What is clear is that some very well respected designs aren't actually that good, and there are very large differences in performance between the good and the bad. I'm currently working on a very high performance local oscillator/multiplier strip for my own use, and I've run simulations using Genesys and Spice (and, in a couple of cases, Serenade) of oscillator designs using a model of a good quality 128MHz 5th overtone crystal (Qu = 64000) and NE856xx transistor(s) (a highly recommended device BTW) in the maintaining amplifier. These are my results after optimisation:

Single transistor Butler oscillator (like K3VOT's): Qu=2400

'Balloon board' Butler oscillator (typically used in many UK microwave designs): Ql=4600

Single transistor self-limiting impedance inverting oscillator: Ql=16000

Cascode Pierce oscillator with separate limiter: Ql=24000

Cascode Driscoll oscillator with separate limiter: Ql=33000

The Driscoll circuit was originally designed for use in 5MHz frequency standards, but it can be made to work very well at VHF. It's capable of extremely good noise performance, while not over dissipating the crystal.

My oscillator strip uses a cascode Driscoll with some additional circuitry to further improve the very close-in noise. It has facilities to phase-lock the VHF oscillator to an external standard. The design uses mainly 0603 parts and has 'no tweak' filtering. Currently the output is in the 400MHz region, but I'm working on multipliers into the microwave region.

Improving Crystal Oscillator Signal Purity

John Hazell, G8ACE

Many microwavers purchase low cost crystals from providers such as Quart-Slab, Klove and Eisch. At 24GHz and above it sometimes happens that the received signal using a new low cost crystal can sound full of LF jitter or perturbations. This makes copying of weaker SSB signals quite difficult.
The blame for this effect can be laid directly at the door of the crystal itself.

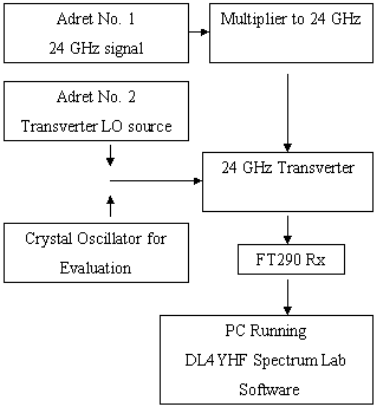
I have been pursuing a cure for this problem for some time, it being quite difficult to get suppliers to discuss the problem and therefore track down a solution. The crystal blank quality on low cost crystals appears to drop to very poor quality at times, possibly quality that is adequate only for computer crystals. Both G4BRK and myself purchased 100.2 MHz crystals from Klove recently and both these showed large amounts of jitter. Other crystals purchased at the same time from Klove are perfectly good but because the frequency was different they where almost certainly made from different quartz blanks of better quality.

So the first unknown is the **crystal blank quality**. It seems likely that once a bad quality crystal frequency appears its likely to persist until all that stock of blanks is exhausted.

Additionally there appears to be a likely hood that variable stresses are contained in the bond wires, the crystal element within the can adding to this jitter problem. QuartSlab suggested some time ago that holding a crystal vertically and using a soldering iron to heat the lead out wire for ten seconds to allow the heat to travel up inside the unit would reduce the jitter.

This was tried and an improvement in the amplitude of the jitter of perhaps 2:1 was obtained. I also understood that professional users often store new crystals at around 80º C until they are required. I had kept some crystals under the jacket of my hot water cylinder for several years but on finally using these the jitter was no different to a new crystal. This warm storage may help with ageing though I have no measurements to support this aspect.
Another crystal supplier then suggested cycling the troublesome crystals between around +80º C and -10º C for one hour at each temperature for 48 hours. A simple arrangement to do this cycling was to use my OCXO[1] unit itself to do the crystal heating. A resistor of 4K7 was added from TP1 on the circuit diagram to the +10.5v regulated supply. This allows the upper oven temperature range to be extended. Additionally if the heating is done using the complete OCXO it is important that high temperature Epoxy was used to bond the heater plate to the PCB board. For the low temperature simply placing the bare OCXO module in the freezer was the simplest solution.
Only the crystal itself needs to be cycled so if you have a means then it can be done outside the OCXO.

An evaluation method needed to be established to verify any improvements that might be made. The **system diagram** (seen here on the following page) shows the arrangement employed. Two Adret 5104 synthesizer units are employed to establish a reasonable quality reference. The Adrets themselves will not be perfect but the results show them to be considerably better than a 'bad' crystal. The IF output from the transverter is connected to the FT290 Rx and the resulting CW audio note connected to the sound card in the PC and the DL4YHF[2]



Spectrum Lab software.

Figure 1, shows the frequency perturbations of the test set-up. Vertical scale is 10Hz/div. Horizontal divisions are 10 seconds. The amplitude of the jitter is around 10-15Hz pp. This amount sounds perfectly satisfactory on CW.

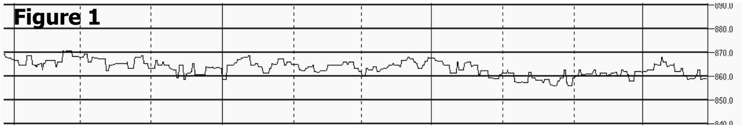


Figure 1

Substituting the test crystal in its OCXO for Adret No. 2 a measurement of that can be made.

Figure 2 shows the new 100.2 MHz crystal as considerably worse than the test setup. Vertical scale is now 25 Hz/div. The jitter amplitude being as high as 300Hz pp. This is very audible as a nasty wobbly note. The method of heating and cooling the crystal is rather arduous and was done as time permitted. The crystal being re-measured at intervals for signs of improvement. Temperature cycling during the daytime over a period of around five days gave the result in **Figure 3**.

Here the vertical scale is 20 Hz/div. So it's possible to see around 30Hz pp. jitter and a result about 2:1 worse than that of the Adret but an overall improvement of about 10:1 for this particular crystal.

Another crystal, for which a graph is not available, improved to match the Adret result so it's possible that the Adret jitter would be the limiting measurement factor for this case.

The improvement is not necessarily as good for all crystals. If you have the jitter problem then only by doing the temperature stressing will you find how good the crystal is capable of becoming. The sharp frequency transitions shown in Fig. 2 are assumed to be stresses in the bonding wires which presumably are

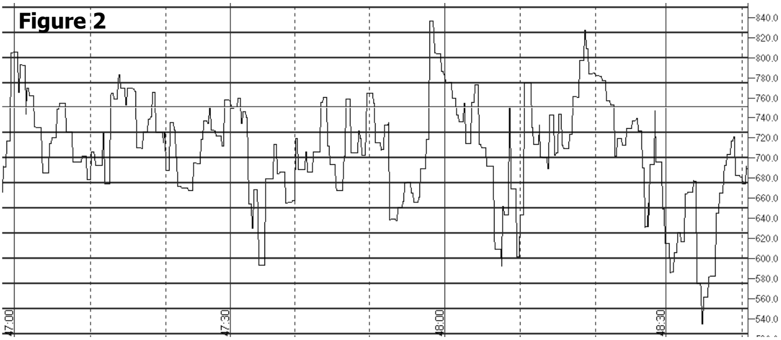


Figure 2

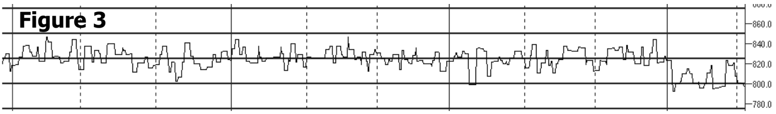


Figure 3

annealed in the temperature cycling process along with some stresses within the crystal resonator.

Luis Cupido, CT1DMK, on a recent visit suggested increasing the crystal drive level. This I assume will help accelerate the shedding of any rubbish on the crystal itself. The design drive level in my OCXO is nominally 750uW to the crystal. By shorting R2, the effect of the amplitude limiter diode D1 is removed and by shorting R8 the crystal drive level is raised to 10mW whilst maintaining around +4dbm output level. A really high power oscillator is possibly needed but there is some evidence that running the OCXO for several days at this higher drive level has some improvement effect. The tests have been done at 24GHz because the writer's portable transverter design allows easy testing. Testing at 10GHz would be possible but the amplitude of the measurements would be around 2.3 times smaller. It is assumed from the crystals evaluated so far that this jitter only manifests itself as an operating problem at 10GHz and higher. Regrettably the degree of improvements that can be obtained can only be found with the extended cycling and re-testing. So far all crystals have improved to a greater or lesser extent.

It has been reported and observed that the problem does improve on its own over a very long period but I assume this is due to the slower action of a crystal being heated and cooled in its application at a much lower cycling rate. If you have a problem crystal then testing and cycling should reward you with an improved LO note purity.

References:

1. http://www.microwaves.dsl.pipex.com/
2. http://www.qsl.net/dl4yhf/spectra1.html

Synthesisers
PLOs and DDSs

Experiences with the San Diego Microwave Group 1152MHz Synthesiser

Peter Day, G3PHO

This article outlines my own experiences with the 1152MHz Qualcom synthesiser board. The board, now used as my primary PORTABLE microwave frequency marker.

The pc board is quite large! It measures some 6 by 9 inches and so requires, by some standards, a bulky enclosure. I housed mine in a plastic box obtained from Maplin. A plastic box was deliberately chosen as the module is used as a multi-microwave band frequency marker when I am out portable. It radiates strong, very stable signals on all bands 2.3GHz to 24GHz (and maybe higher). It actually includes a great deal more circuitry than the basic synthesiser — a 1GHz IF amplifier and receive/transmit amplifiers for example. Originally, the VCO operates on a frequency of 750MHz to 1000MHz but a simple modification easily gets it onto 1152MHz. A high quality, 10MHz TCXO reference oscillator is divided down for the internal PLL reference frequency of 1MHz. The synthesiser chip is a 40 pin Qualcom 30361-16N or 32161-16N.

If one of the onboard MMICs is enabled as an amplifier, some +10dBm RF output at 1152MHz is available … an excellent LO source, especially for a beacon. In my case I did not do this modification, preferring instead to have the rich harmonic output which read 4mW on my power meter (probably at a multitude of frequencies at the same time!).

There are five simple modifications to be made to the board to get it to operate as a microwave marker. These are detailed on notes provided with the pcb. One of the modifications, adjusting the 10MHz TCXO, was not possible with my board as the 10MHz module was in a sealed can and was marked as having zero kHz offset. In reality it turned out to have a slight offset at 4.8Hz at 10MHz resulting in the 10GHz marker being at 10368.005MHz. However this is no real problem as the frequency is very stable, providing a constant, accurate marker just inside the narrowband portion of the band. On 24GHz it is just 12kHz up from the bottom of the band, no big deal with an IC202 as the RX IF!

Most of boards obtained from Chuck have an adjustable TCXO so purists can get theirs spot on all the zeros! Unlike other frequency sources I have, no multiplier diode was required at the RF output connector to produce harmonics above 1152. The module radiates plenty of those with no help at all! In fact the connector is left unloaded. When a Sat TV multiplier diode was connected at this point (in the manner described in the Newsletter last year) the output broke up into an unstable mess! One minor disadvantage with the module is the power supply requirement. Two voltages are need…. +15V DC and +5V DC. The latter draws around 0.5 amps and is thus a potential battery "zapper" when out portable over long periods. The +15V supply is only required to produce around 60mA or so. It actually feeds a 12 volt regulator which then supplies the 10MHz reference oscillator and so it would be possible to run the whole board from a 12V battery and use a 3 pin, step-down regulator (i.e. a 7805) for the +5V circuitry.

In my case I decided to keep the 15V input and built up a small, external 12V DC input power supply that produces the +15V from an LM2577T-Adj "switcher" IC (see Microwave Newsletter, May 1997) and the +5V from a 7805 regulator. The latter requires good heat sinking but the switcher IC runs very cool with no heat sink. A small die cast box houses the PSU. Leads in and out of it are thoroughly filtered to prevent any radiation from the switch mode section. The heavy current consumption at 5V was solved by having the inputs to each IC separately switched. Provided the 15V supply is left running, no drift occurs if the 5V regulator is switched off until a marker signal is required. Instant lock is achieved on switch on from cold. A surface-mount LED on the Qualcom board just flicks on for a second and then goes out, indicating lock.

Unfortunately there is no circuit diagram available with the board, as received from San Diego, so it was not possible to decide what to cut out of the board that was of no use, in an effort to reduce overall current consumption. Under portable conditions, I switch on the 15V supply to the TCXO just before I set off to the /P location. By the time I am there (usually 30 minutes) the module is drift-free. Once the microwave receiving equipment has settled down I then switch on the +5V supply and the marker is instantly available as an accurate frequency reference, whenever I need it. Prior to using this system I had an old Microwave Module 384MHz, 500mW source (suitable re crystalled) driving a G3JVL type multiplier using MD4901 varactor. I had already fixed a Murata crystal heater into the MM module but even then the stability over, say, 10 hours was not really good enough for me to be confident of "spotting" frequencies on 10GHz within +/- 1kHz accuracy. The Qualcom module can sit on the car dashboard, away from the vagaries of the weather outside and thus remain on frequency throughout the day's activities. Short of taking out my precious 90-120MHz Adret synthesiser or locking my transverter LO to a TV-locked or MSF-locked source, I cannot think of a more accurate way of keeping on frequency under portable conditions. Of course, the situation at home, where a good frequency counter and the Adret are on 24 hours a day, 365 days a year, is completely different! The Qualcom 1152MHz synthesiser is therefore recommended as a valuable tool for portable microwave work.

Oscillator Sources for Microwave Use

Andy Talbot, G4JNT

For some time I have being looking at ways of locking microwave source oscillators to a master frequency standard to allow high stability operation for propagation monitoring of beacons and to remove frequency drift with temperature and ageing. Most good quality frequency standards operate at either 10MHz or 5MHz, whereas microwave sources generally start from a crystal in the 100MHz region and are multiplied up. Going from 10MHz to some arbitrary 100MHz value is not straightforward. For local oscillators which generally operate at multiples of exact MHz, it is sometimes possible to find frequencies that can be generated relatively easily from 10MHz (e.g. 106.5MHz for a 10GHz LO source when multiplied by 96) but a beacon on 10368.905 (GB3SCX) needs 108.0094270833MHz - not nearly so easy!

The Conventional Approach

The approach taken by WA6CGR (www.ham-radio.com/wa6cgr) in his phase locked microwave source is to use a Phase Locked Loop with arbitrary values of division for both the reference and VCO divider chains, as shown in Figure 1. The output frequency is given by :

$$F_{rf} = M * N / R * F_{reference}$$

where M is the RF multiplication, N is the PLL VCO division and R is the reference division value. Is usually possible to find some pair of values of N and R where the resulting locked frequency is "near enough" to the wanted value to be acceptable. N and R should not be made too high as otherwise the divided down comparison frequency becomes unmanageably low, and some arbitrary value has to be decided on when working out R and N values, which usually have to be derived by trial and error, or more easily by a computer search of all possible values, given the dictates.

For the example above comparison frequency was specified as being 10kHz minimum (see below for justification) and a computer search came up with values of R = 849 and N = 9170 with an RF multiplication M of 96 for a result which is 406Hz low - probably good enough in practice. The comparison frequency ended up at around 11.79kHz. For our purposes on microwaves, we probably have one of the most stringent stability and phase noise specifications of any user on these frequency bands. Microwave amateurs are (probably) the only group who are concerned with phase noise performance at GHz that is only tens or even hundreds of Hz away from the carrier centre. Poor phase noise performance in this region manifests itself as an annoying rough tone to recovered SSB or CW signals, and even closer in to the carrier as a frequency jitter or wandering. So the basic microwave oscillator source at 100MHz has to have a very good performance so that when multiplied up many times its own phase noise (which is multiplied by the square of the multiplication factor) is acceptably low. 100MHz overtone crystal oscillators are usually quite adequate in this respect, although their frequency setting accuracy, temperature drift and ageing leave a lot to be desired. Furthermore, unless they are designed and cut for operation at elevated temperature, putting them in an oven or adding a clip on crystal heater hardly helps in temperature stability which can cause several parts per million change even over day / night inside a house. A varicap diode in series with the crystal can easily be added to and allows the crystal frequency to be pulled a few parts per million - a few hundred Hz - by varying the control voltage applied to the varicap. This item will be referred to as the VCXO, or Voltage Controlled Crystal Oscillator.

10MHz reference oscillators on their own, free running, make use of the very best quality crystals there are, and the inherent phase noise of these if they could be multiplied up directly is usually more than good enough for our final requirements. However, being forced to use a phase locked loop in the frequency determining process can undo all the oscillator manufacturers good work! So if we can phase lock the crystal to a divided and multiplied version of a reference we're home and dry. The main problem here is the restricted pulling range of the 100MHz overtone crystal oscillator. As a rule of thumb, the bandwidth of a phase locked loop - which dictates the time required to lock up and the phase noise performance - has an absolute maximum value given by the product of the Voltage Controlled Oscillator control constant (Kv) defined in Hz/Volt and the Phase Detector constant (Kd) defined in Volts per radian. Where the output of the VCO is divided down, the Kv value has to be divided by N.

To a first approximation the absolute maximum PLL bandwidth is given by:

$$BW = Kv . Kd / N$$

Phase Locked Loops in other applications often filter to narrower bandwidths, but for synthesisers we always want the absolute maximum bandwidth that can be achieved to minimise any noise added onto the VCO from external or internal sources.- this can make PLL design both easier - in the area of loop filtering, and also more difficult - in the area of high speed phase comparator performance. For the GB3SCX example above, Kv of a typical VCXO Butler oscillator was measured at 250Hz/Volt (divided by 9170), and a standard edge triggered phase detector in a 5V system usually has Kd = 1.6 V/radian, giving a maximum loop bandwidth:

$BW_{(Max)} = 250 \times 1.6 / 9170 = 0.04$Hz or 1 cycle per 23 seconds

Since loop lock and stabilisation from switch on up can take several cycles we are looking at a period of several minutes before an acceptable output stability is achieved. Furthermore, and even more serious for our purposes, the inherent stability of the VCXO has to be stable in its own right within this bandwidth period, as otherwise the loop cannot correct it. Asking an non-ovened crystal oscillator to hold a few parts in 10^{-8} several tens of Hz at the final 10GHz microwave frequency) within even one cycle of loop bandwidth is not really on with temperature shifts and draughts. So it looks as if even the VCXO has to be oven controlled itself if taking this route. I first encountered this VCXO / PLL problem not originally on the microwave bands, but at LF when playing with ultra narrow band processing where the frequency stability requirements, (in relative terms anyway) are similar. A 24.576MHz clock oscillator had to be locked to a 5MHz reference and the only common frequency for comparison was 8kHz. It was necessary to allow 30 minutes for satisfactory PLL stabilisation and this was using a fundamental mode crystal with inherently higher pulling ability.

Direct Digital Synthesis

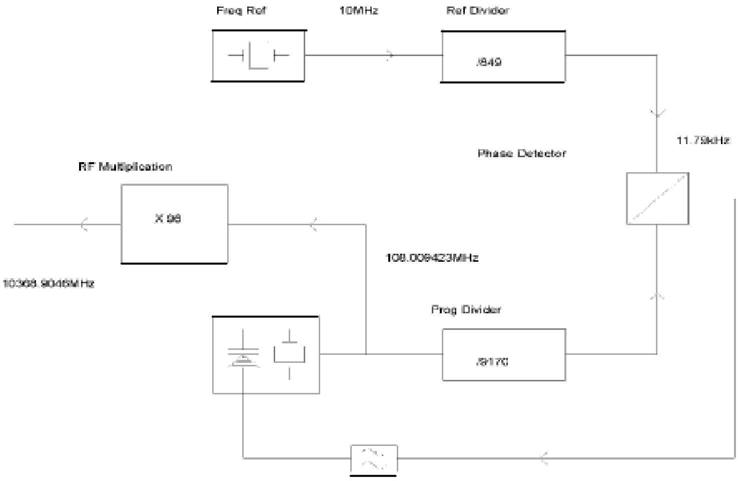
A few years ago DDS devices became widely available to amateurs and it became easy and straightforward to construct a source of virtually any arbitrary frequency up to a value of about 40% of the clock applied to the DDS -

which is typically in the 100 to 200MHz region for low cost devices. See my DDS module design published in RadCom November 2000. This used an AD9850 DDS with a maximum clock of 120MHz, allowing frequencies up to around 40MHz to be generated with a resolution, or tuning step, of $F_{clock} / 2^{32}$. A later device, the AD9851, which is allowed a clock frequency up to 180MHz, and also had an internal X6 PLL multiplier on board the chip itself so the actual clock source need only be 20MHz maximum. The beauty of DDS solutions is that there is no inherent increase in phase noise due to the DDS process - phase noise in is scaled exactly by the frequency division process.

I did wonder if the internal X6 in the AD9851 added any phase noise, but it is a very high bandwidth PLL and the data sheet states that his is "insignificant". Tests by multiplying a DDS output up to 10GHz did tend to agree with this statement. By driving an AD9851 with 10MHz from a frequency reference, a 60MHz clock results which can generate any frequency from 0 to around 24MHz in steps of 0.014Hz The downside to DDS sources is that they are not very clean, 50 - 60dBc spurii are common, and these seem to bear no relationship to the carrier frequency (they do, but the relationship is complex and is based on harmonics, alias products, harmonics of alias products and intermods). Suffice to say, the sprogs are troublesome, and some low level ‘sprogs’ are quite close to the carrier, albeit many dB down. I initially thought the DDS output at 21.6MHz could be multiplied by 5 directly to 108, and did try a breadboard of this, generating a test signal multiplied up to 10368GHz. It sounded terrible!

The centre frequency has a 'pure' enough tone, but it was surrounded by whiskers and rubbish that sounded like very high level of phase noise. What was happening was that all the close in spurii, that may well have been 80dB down and undetectable on the source, when multiplied by 540 were increased by the square of the multiplication factor and were now very significant. This idea was very quickly abandoned. **Next I considered the PLL route - see Figure 2.** Since the DDS can generate tens of MHz directly, the loop bandwidth could be made very high—certainly in the tens of kHz region and ought to be capable of stabilising a poorer quality oscillator such as an LC design. Various attempt were made to do this, but a sufficiently high bandwidth PLL never materialised. Edge triggered phase/frequency detectors require high speed logic to make them operate and I was losing patience at this point, quite apart from the fact that the DDS spurii were still proving difficult to filter out with loop bandwidths sufficient to clean up an LC source. More perseverance with good LC oscillators and fast logic may have come up with something, but I could be bothered.

So, back to a crystal oscillator as the source - at least these are always clean. Now a PLL using this as the VCXO does not have to divide down by a large factor, as the comparison can be made at tens of MHz. By using a simple divide by 8 prescaler the VCXO can be brought down to 13.5MHz for comparison with a reference generated from the DDS. The exact division ratio is unimportant as the DDS can generate any frequency, and a divide by 10 would work just as well. A simple mixer type phase detector was implemented using an NE612 receiver chip followed by a low noise op-amp, rather than trying to make a complex high speed edge triggered D-type flip flop design. Kv of the VCXO was 250Hz/V now only divided by 8, and the NE612 / amplifier combination had a Kd of 2 V/rad - or more depending on the amplifier gain.



Now, maximum loop bandwidth = 250 * 2 / 8 = 63Hz ….. much more acceptable.

The circuit of Figure 3 was built up and tested. It locks up virtually immediately, the note when multiplied up to 10GHz sounds 'perfect' and all appears to work as required. No particular effort was made to optimise loop filtering - as the inherent loop bandwidth is dictated solely by the Kv.Kd product this was just left to cope as best it could and give the highest possible bandwidth, with no attempt to control overshoot or anything. As with any PLL based on a mixer rather than an edge triggered phase detector, the loop will only lock up if the initial frequency error at switch on is within the loop bandwidth. So the crystal oscillator has to be within 63*8 or around 500Hz of the wanted frequency. In practice, and to cope with temperature shifts, a starting frequency of significantly less than this should be aimed for - say within 200Hz which corresponds to a couple of PPM. This is getting a bit tight for an non-ovened crystal to be used over a wide frequency range, but, with decent overtone devices, it can usually be achieved. The divide by 8 prescaler chip used here, the SP4908 is probably obsolete now, but I had one Any suitable device from the junk box will do, and doesn’t have to divide by exactly 8 - so long as the final frequency is within the range of the DDS source. VHF prescalers are not too common now as synthesiser chips themselves operate at 1 - 2GHz; the prescalers that are around tend to operate at many GHz, but even one of these can be pressed into service as they usually operate down to DC. An incidental advantage of the loop locking up quickly it that is just about possible to generate frequency shift keying by reprogramming the DDS. Provided the frequency shift is small - no more than around 1 to 2kHz - the loop can track this change within a few milliseconds. If CW at around 12WPM is used then the keying just sounds a little soft, as the tone changes from mark to space. Much faster than this, and the edges are smeared out too much.

The DDS frequency resolution is $F_{clock} / 2^{32}$ which for a 60MHz clock is 0.014Hz. When multiplied up to 10GHz by the X8 prescaler and X96 RF Multiplication, this step size translates to a little over 10.7Hz - probably accurate

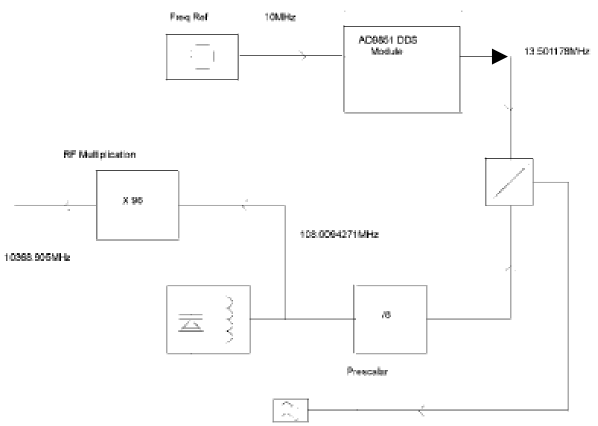


Fig 2 DDS Source for generating the PLL reference

enough for most purposes!

An afterthought that occurred to me as this article was being written ...

Instead of using a prescaler to divide the VCXO down to a frequency that could be generated by the DDS, why not use the VCXO directly as the DDS clock (without the X6 PLL option enabled)? Then by thinking backwards, programme a value into the DDS that generates exactly 10MHz from the chosen oscillator frequency. As the internal PLL is not needed, and the device is only operating at 100MHz the original AD9850 can be used in this adaptation. This concept will form the basis of the Mark two design. Generating the accurate 10MHz signal is a subject I’m also closely involved with, but that will have to wait for another article.

Another possibility for the next generation is using the AD9852 DDS. This chip can have a clock of up to 300MHz and an internal PLL which is programmable between 4 and 2. It also has a 48 bit accumulator so frequencies can be set to a resolution of 2^{-48} of the clock. At 108MHz clock frequency, followed by multiplication to 47GHz, this still allows a final frequency resolution of 0.16mHz (no, that is **not** Megahertz, it’s **millihertz** !!!)

More afterthoughts ...

I'm not sure GPS LOCKING is necessarily the best route for high frequencies - with the long time constants needed, when used in /P or situations when the temp changes the freq can wander around quite a lot whilst still being "locked". I'd rather maintain a very good 10MHz source in a separate box with its own batteries that can be carried around and calibrated when necessary.

Locking is good idea, though, for beacons and home stations where a reference can be maintained always switched on - but still with battery backup for coping with power cuts. I much prefer using GPS to gate a frequency counter that then measures the reference oscillator freq over a long period - 1000 or 10000 seconds (17 minutes or 2.8 hours) for calibration only. Displaying 5 or 10MHz on a scope triggered at 1 pulse per second is feasible for getting very accurate calibration, provided you do it in a darkened room.

For the best of both worlds, make a locked source for home use only (the MSF one is probably good enough), then check the portable unit against this using a scope which will only take a few minutes to set both traces to appear stationary against each other.

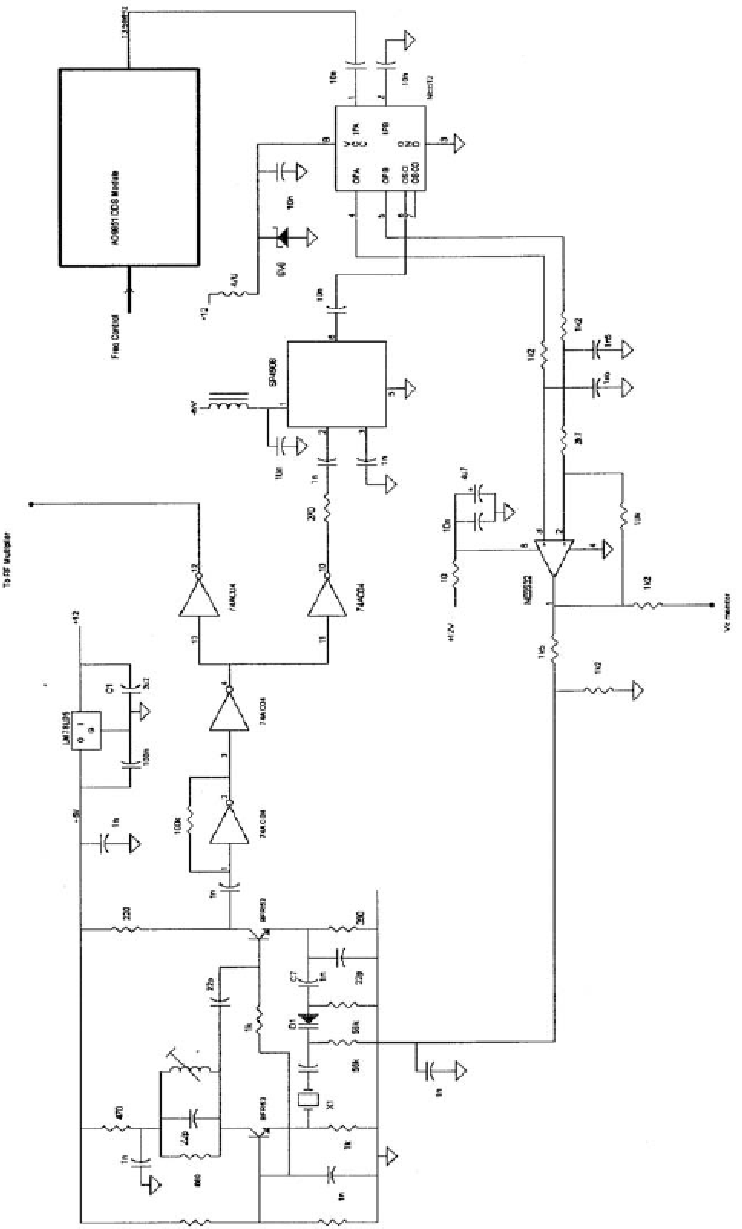


Fig 3 Circuit of DDS Locked Microwave Source Oscillator

A simple way of phase locking microwave local oscillators

Andy Talbot, G4JNT

In the article on 'Locked Oscillator Sources for Microwave Use' published in the last Microwave Newsletter, I made the throwaway comment, in the first paragraph, that phase locking local oscillators with nice round frequencies of integer MHz was simpler than producing an arbitrary value for beacons, etc. Well, after saying that I had to come up with something and here are a few notes on a simple breadboard method of phase locking these crystal oscillators to a reference source. The circuit is presented as a starting point - there will be some experimentation needed for each individual case.

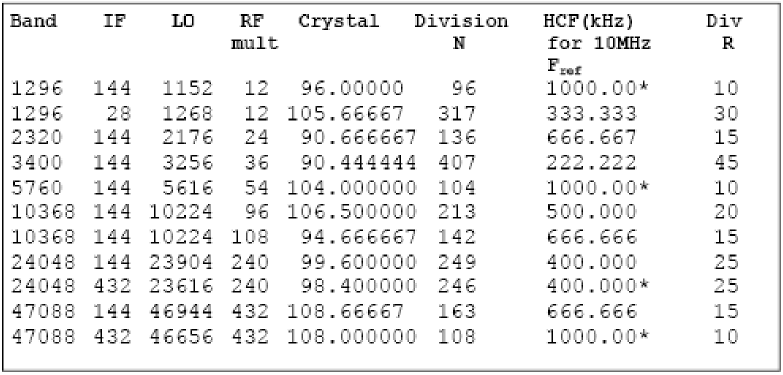
For the popular microwave bands the LOs are usually generated from overtone crystal oscillators followed by multipliers, with the following table showing the usual LO frequencies for the narrowband segments along with the associated crystal frequency. The final columns show the highest frequency that is a sub multiple of both this and a 10MHz reference, the highest common factor or HCF, and the associated division. This HCF can become the comparison frequency in a phase locked loop, and is the highest frequency that is possible here.

Two things become obvious:

All the comparison frequencies can be derived from a 10MHz reference by making use of simple logic divider chips to give the divide by R function (all could be derived from 2MHz in fact) All the comparison frequencies are over 200kHz, so phase locked loops can be made with wide loop bandwidths. For those marked with a *, an even higher comparison frequency is possible, but the values stated keep the frequencies within a narrower band for a common design.

So now the only difficulty is providing the divide by N from the crystal frequency. An off the shelf synthesiser chip such as the MC145170, or those from other manufacturers, would make an easy job of this but there is an even simpler solution providing you are prepared to do a bit more adjustment and optimisation.

Anyone who has studied the 'brick' range of microwave sources will have seen how a high Q cavity oscillator is locked to a reference oscillator in the



Band	IF	LO	RF mult	Crystal	Division N	HCF(kHz) for 10MHz F_{ref}	Div R
1296	144	1152	12	96.00000	96	1000.00*	10
1296	28	1268	12	105.66667	317	333.333	30
2320	144	2176	24	90.666667	136	666.667	15
3400	144	3256	36	90.444444	407	222.222	45
5760	144	5616	54	104.000000	104	1000.00*	10
10368	144	10224	96	106.500000	213	500.000	20
10368	144	10224	108	94.666667	142	666.666	15
24048	144	23904	240	99.600000	249	400.000	25
24048	432	23616	240	98.400000	246	400.000*	25
47088	144	46944	432	108.66667	163	666.666	15
47088	432	46656	432	108.000000	108	1000.00*	10

100MHz region by a sampling phase detector. This device combines the functions of frequency multiplier and phase in one network. In the bricks, a snap varactor diode is hit with about 200mW of reference signal and so generates sharp sub-nanosecond pulses at this rate. These pulses are applied across a pair of microwave diodes, forming one input to a single balanced mixer whose other RF ports is fed with a portion of the cavity oscillator signal picked off via a small probe. This sampling mixer approach to PLL design makes for considerable simplicity as it inherently removes any need for a high frequency divide by N circuit - but has two major drawbacks. Firstly, the VCO can lock to ANY harmonic of the reference, and in the brick designs this is prevented by restricting the electrical tuning range of the VCO to less than half of the comparison frequency's Sometimes a pull in range of only 5-10MHz (at the fundamental L-Band frequency) can be observed.

Secondly, the output from the phase detector is at a very low level - typically a few tens of millivolts per radian rather than the 1.6V/radian of normal logic. In effect, the complete voltage range that would have been possible for the drive power is having to be shared over every one of the comb frequencies. However, a low noise op- amp can easily provide DC gains of the hundred or so required here, and differential amp doing this job inside the bricks is clearly visible when the side cover is removed.

So, lets try this idea at lower frequencies to lock a VCXO - the very restricted tuning range of crystal oscillators means that drawback 1 is not an issue. Look at the circuit of Figure 1, a two chip R divider (programmable for any value from 1 to 256) generates the reference. This is applied to an impulse generator using the propagation delay inherent in three high speed logic gates plus an additional capacitor to generate a series of negative going impulses of a few nanoseconds in width. If you look at the output at this point on a spectrum analyser, the spectrum will show the classic sin(x)/x shape with a null corresponding to the pulse width. The extra 100pF capacitor in the delay can be adjusted to ensure this null does not fall at the wanted crystal frequency - it can get a bit unpleasantly close with 74AC series gates in this position.

These impulses are at a level sufficient to directly drive a diode ring mixer, so all that is now necessary is to apply a portion of the VCXO to the other mixer port, amplify and filter the IF output and feed back to the VCXO for a complete PLL. As in all the microwave sources I've discussed so far, the PLL needs to have as wide a bandwidth as possible to remove VCXO close in phase noise and jitter, so there is not much effort that has to go into filtering, apart from removing the fundamental comparison frequency component. This last point is significant, don't go too low with the reference frequency as filtering it out will be more difficult, as well as the fact that the voltage swing available from the diode ring will be even less. The pair of 74HC161 devices forming the R divider could be replaced by appropriate sections of a 74HC390 chip, or an 'HC90 with feedback in some cases, but this solution gives a generalised divider allowing factors for R divisions not in the table, such as those with 7, 11, or 27 in them that couldn't be obtained from a simple configuration. It works be preloading in the number set on the links, then counting up to 256 where it overflows and loads in the preset value again repeating the process. So, the wire links have to be configured to load in a value of (256 - R). With HC CMOS, a logic '1' must not be generated by leaving an input as an open circuit, as could be done in the old days of TTL (and even then it was unethical!), so each input must be tied to either +5V or ground.

Results:

My breadboard was not exactly as shown in Figure 1. Instead, I started off with a 1MHz signal rather than 10MHz and divided by 3 in a single chip to give 333.333kHz which was used to lock the 94.666MHz crystal in my 'JVL based 10GHz system. Figure 1 has been built, and just committed to PCB, but not yet into any finalised system. The first version for 10GHz worked fine, except for the fact that the VCXO drive level to the mixer was rather critical - this has to be kept at about 1mW for the mixer to remain linear, whilst maximising the DC output level from the IF.

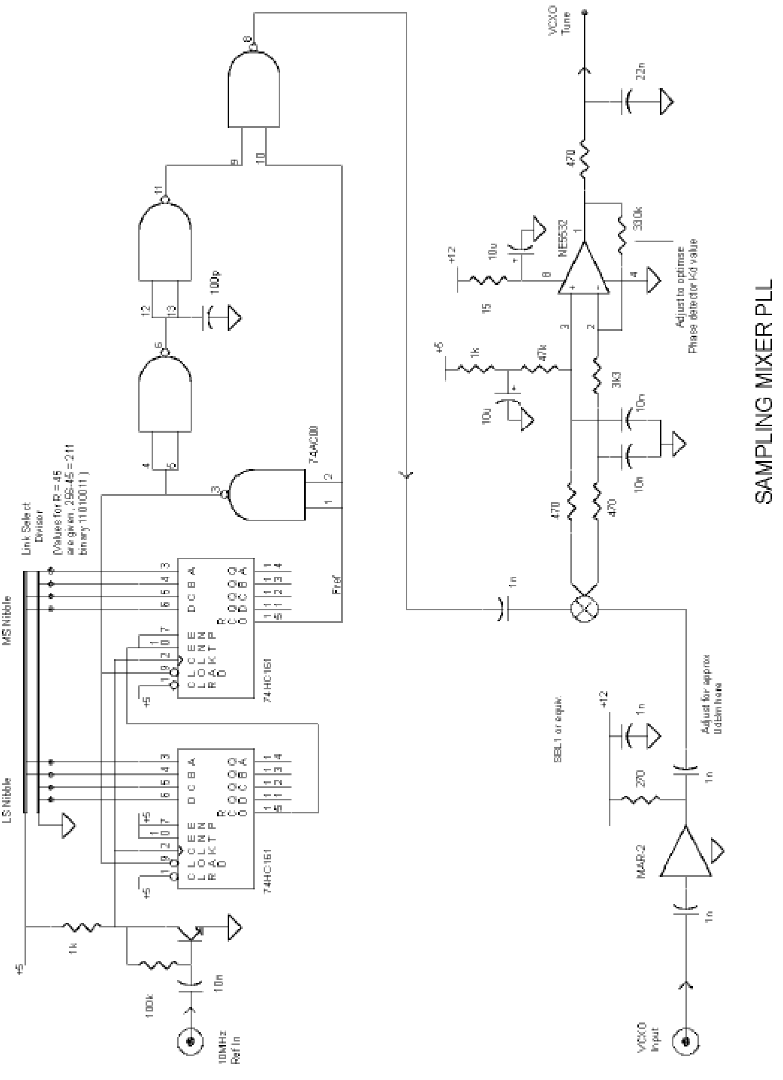
If drive level is increased too much, the mixer saturates and, surprisingly, IF output falls off in this usage. I'm not too sure why this should happen, but is probably due to the short pulse forming the pseudo-LO, and its very low means value. Real experimenters may like to try replacing the packaged DBM with a pair of diodes plus transformer/balun for a single balanced design as in the bricks. This may give more sensitivity and less dependence on RF drive level, but I couldn't be bothered with this level of fine tuning once the DBM had proved itself.

Note that the mixer also has to be one where both sides of the IF port are accessible. Most devices like the SRA-1 and SBL-1 offer this, but higher frequency mixers sometimes ground one side of the IF so in this event the op-amp will need to cope with a negative input voltage. The DC gain in the OP-amp circuitry is what was required for this 333kHz reference - other higher comparison frequencies will allow proportionately lower gains. The exact circuit configuration is not a completely true differential amplifier as this is not essential, it also has to transfer the 5V reference through to the output. As true differential operation is not necessary, to change the gain it is only really necessary to alter the single resistor shown as 330k in Figure 1.

An easy way to set up the necessary DC gain is to look at the op-amp output on a scope with both reference and VCXO signal applied, but with the input to the VCXO tuning diode clamped to the 5V centre value. Then by manually tuning the VCXO through its range a beat at the difference frequency will be seen, this will be centred on exactly 5 Volts and its peak to peak value must be wide enough to be able to tune the VCXO at all ranges of temperature and drift …. a value of 2 - 5 volts peak-to-peak will probably suffice for most VCXO designs. The op amp gain may need adjusting to get to this figure, and the value also depends to some extent on the RF level to the mixer.

With no reference applied, the VCXO needs to be adjusted to the wanted frequency with 5V on its tuning line …. try to get as close as possible as the tuning range is not very wide once temperature and crystal ageing are taken into account. My breadboard locks up near instantly when connecting the reference, provided the crystal heater has been pre-warmed! If a crystal heater is used, the initial cold start frequency will probably be outside the PLL lock range; in my case it takes about 10 seconds for warm-up before the PLL will lock from a cold start.

Now, the only drawback to getting within 0.1Hz on my 10GHz system (and yes, I can get 10-11 on the reference) is the accuracy of the 144MHz IF. The IC202 LO is derived from a DDS (more about this another day) controlled from a bog-standard packaged 100MHz crystal oscillator and rotary up down counter plus LCD display to 10Hz resolution and 100Hz readout. The DDS could have been driven from the 10MHz reference multiplied up, but there is still the inherent accuracy of the carrier crystal inside the IC202, so I can only get to within tens of Hz of the wanted frequency at 144MHz without doing some pre-calibration of the IF, and even then it will drift a few more Hz over the day. When I designed the digital IC202 LO I hadn't anticipated wanting a better accuracy for microwaves than this, but some recent tests between G8ACE and G3NNS make this assumption invalid now.


SAMPLING MIXER PLL
G4JNT 2004/04/07
10MHz Ref In
100k
10n
1k
+5
LS Nibble
MS Nibble
Link Select Divisor
(Values for R = 45 are given, 256-45 = 211 binary 11010011)
74HC161
74HC161
Fref
74AC00
100p
VCXO Input
1n
MAR-2
270
1n
+12
1n
SBL1 or equiv.
Adjust for approx 0dBm here
1n
470
470
10n
10n
3k3
10u
1k
47k
+5
15
10u
+12
NE5532
330k
Adjust to optimise Phase detector Kd value
470
22n
VCXO Tune

DDS Module Controller Software

Andy Talbot, G4JNT

Overview

The DDS module is supplied with a PIC microcontroller that contains code to allow the AD9852 to be controlled via an RS232 serial link from an ASCII terminal such as a PC running Hyperterm software. A further input to the PIC can be used for an external trigger, so the updating of the DDS output can be synchronised to an external even such as a UTC pulse from a GPS receiver.

Setting up the serial link

Set your terminal programme to 19200 baud, 8 data bits, no parity and 1 stop bit (19200 N81). Turn off local echo and do not enable CR-LF translation on receive. Connections between the PC and the DDS module are defined in Table 1:

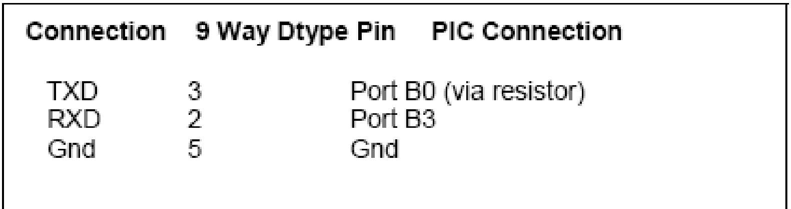
Connect the serial link and switch on the DDS module. After about half a second delay, a display similar to that in Table 2 will appear.

All the registers in the AD9852 chip are loaded with the values stored in EEPROM, either the default initial values, or any that have been subsequently changed by the G command (see below). As supplied the default will result in an RF output at exactly 0.25 times the clock frequency, with no PLL multiplier in use. All other

Controller Software

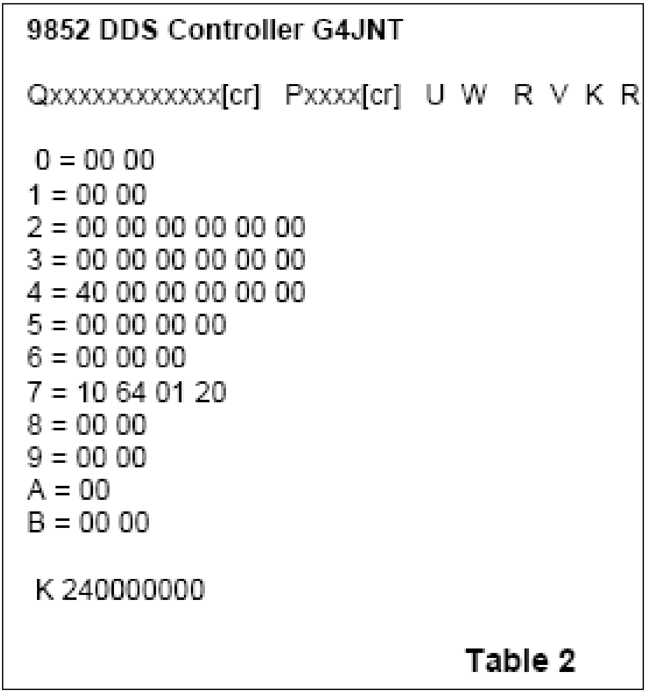
The AD9852 DDS chip has got quite a comprehensive set of capabilities which are controlled by writing appropriate values to its working registers then triggering (updating) the device. The controller software has two main functional capabilities.

For simple frequency generation and phase shifting requirements, a straightforward command structure for quickly updating CW frequency or phase is implemented, with separate commands to update the DDS with these changes either immediately, or on an external pulse edge. The option of writing the current frequency to non-volatile storage (internal EEPROM) for immediate start up next time the module is turned on is also possible. The other mode of operation allows any of the internal registers to be written individually, giving full access to the chip's functionality. All register contents altered using these commands are stored in EEPROM and loaded in the next time the module is turned on. A final option is available to allow users to store a string of up to



Connection	9 Way Dtype Pin	PIC Connection
TXD	3	Port B0 (via resistor)
RXD	2	Port B3
Gnd	5	Gnd

Table 1: RS232 interface connections



```
9852 DDS Controller G4JNT

Qxxxxxxxxxxxx[cr]  Pxxxx[cr]  U W  R V K R

 0 = 00 00
1 = 00 00
2 = 00 00 00 00 00 00
3 = 00 00 00 00 00 00
4 = 40 00 00 00 00 00
5 = 00 00 00 00
6 = 00 00 00
7 = 10 64 01 20
8 = 00 00
9 = 00 00
A = 00
B = 00 00

 K 240000000
```

Table 2

15 characters in EEPROM for reading back on the serial link; they perform no action on the DDS chip itself. This can be used, for example, to store the clock frequency so any DDS driver software can read this value back, and so be used with several different modules, each with their own clocks.

Commands

Immediate programming

Pxxxx[cr] Sets the phase of the RF carrier output. The format must be exactly as shown with xxxx replaced by hexadecimal ASCII characters. e.g. P8000[cr]. The new phase is programmed into the AD9852 (although not into EEPROM) but does not take effect immediately. The characters are not echoed back to the terminal, so when typing in by hand this has to be done blind. If the command is recognised, the controller responds with a single P followed by [cr][lf].This command controls the contents of DDS register 0 **Qxxxxxxxxxxxx[cr]** Sets a new frequency of the single carrier output. The format must be exactly as shown with xxxxxxxxxxxx replaced by hexadecimal ASCII characters. e.g. Q0280000000AF[cr]. The new frequency is programmed into the AD9852 (although not into EEPROM) but does not take effect immediately. If the command is recognised, the controller responds with a single Q[cr][lf]. This command controls the contents of DDS register 2

U This command updates the DDS chip with the new P or Q values set in the above commands. When complete, the controller responds by sending Z[cr][lf] No carriage return is needed after any single letter commands.

T Triggers the controller to wait for a positive edge on the external timing input (Port B2) before updating the values from the P or Q commands. 'Z'[cr][lf] is returned when the update is done. Note that while waiting for the positive timing edge, the controller will be deaf to any further serial commands and may have appeared to 'hang'. This situation has to be checked for in any driver software by looking for the acknowledgement Z before issuing any further commands.

W Writes the current frequency to EEPROM. This command will only be accepted immediately after a P command has initially been issued. If accepted, the controller responds with a 'Z'

Register Programming

The individual AD9852 registers can be updated one at a time and the values are always stored to EEPROM for immediate start up. The AD9852's registers have different lengths depending on their function, from one to six bytes in length. There is plenty of scope for incorrect operation and unexpected results, particularly when programming the control register. Read the data sheet carefully before changing registers!

The G command is used to update an individual register: functionality. All register contents altered using these commands are stored in EEPROM and loaded in the next time the module is turned on.

A final option is available to allow users to store a string of up to 15 characters in EEPROM for reading back on the serial link; they perform no action on the DDS chip itself. This can be used, for example, to store the clock frequency so any DDS driver software can read this value back, and so be used with several different modules, each with their own clocks.

Commands

Immediate programming

Pxxxx[cr] Sets the phase of the RF carrier output. The format must be exactly as shown with xxxx replaced by hexadecimal ASCII characters. e.g. P8000[cr]. The new phase is programmed into the AD9852 (although not into EEPROM) but does not take effect immediately.

The characters are not echoed back to the terminal, so when typing in by hand this has to be done blind. If the command is recognised, the controller responds with a single P followed by [cr][lf].This command controls the contents of DDS register 0

Qxxxxxxxxxxxx[cr] Sets a new frequency of the single carrier output. The format must be exactly as shown with xxxxxxxxxxxx replaced by hexadecimal ASCII characters. e.g. Q0280000000AF[cr]. The new frequency is programmed into the AD9852 (although not into EEPROM) but does not take effect immediately.

If the command is recognised, the controller responds with a single Q[cr][lf]. This command controls the contents of DDS register 2

U This command updates the DDS chip with the new P or Q values set in the above commands. When complete, the controller responds by sending Z[cr][lf] No carriage return is needed after any single letter commands

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Register Programming

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The G command is used to update an individual register: Type G and the controller responds with ‘Reg No’ No [cr] is needed after the G

Enter a value from 0 to 9, A or B. There is no need to enter a carriage return. The controller responds with ‘x bytes of Reg. Data’ , where x is a number from 1 to 6. Enter the data in hexadecimal (with no [cr]), and as soon as the final character of the requisite number is entered, the controller will respond with ‘Z’. (Note, one byte of data requires two characters, 6 bytes requires 12 characters).

The new value is immediately written to both the AD9852 register and the appropriate EEPROM register The V command is used to dump the entire EEPROM contents, in the format shown in the central portion of Table 2.

User Data

These two commands allow up to 15 characters of user data to be stored and read back from EEPROM.

The K command allows data to be entered. Type K and the controller responds with ‘Enter < 15 chars. of user data’ Enter the characters required, followed by a [cr] to terminate. Any ASCII character is accepted, and all letters are converted to upper case. If an attempt to enter more than 15 characters is made, the controller responds with ‘Overflow’ and the first 15 entered are accepted. In the interests of conformity, if this data is used for clock frequency it is suggested this takes the form of , for example, ‘240000000.00Hz’ so the readout is meaningful and can be easily read and interpreted by driver software.

The R command reads back the user data Issue R (with no carriage return), and the controller responds with the stored string

Frequency markers and references

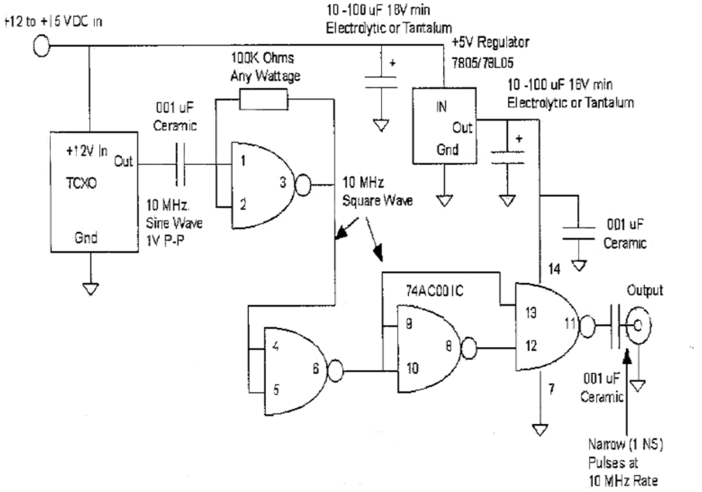
Versatile Microwave Marker Generator

The circuit shown was designed by Kerry Banks, K6IZW, and previously published in the San Bernardino Microwave Society newsletter, which we gratefully acknowledge as the source of information. It is a very simple comb generator that produces low level marker signals every 10MHz, from VHF through microwave frequencies. It could be useful for experimenting and when testing equipment, providing reference points at regular intervals and at a rate more frequent than the normal references which are often based on a starter frequency of 1152MHz.

The 10MHz TCXO in the original design was a surplus Qualcomm item. These devices were available at low cost from Chuck Houghton, WB6IGB by emailing him at:
clhough@pacbell.net
The TCXO drives the circuit, producing 1 nanosecond pulses at 10MHz intervals. If the output is amplified in a simple external broadband MMIC amplifier, you could observe a "comb" of signals with a spectrum analyser. Pass this spectrum through a filter and observe the output on the analyser to assess the filter's passband.

In the circuit shown please note the following:

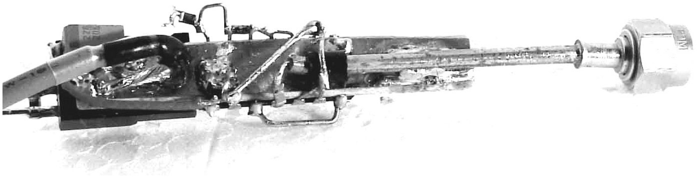
- The Qualcomm TCXO requires +12V but may be substituted with a lower stability TTL Clock Oscillator if the supply voltage is run off +5V.
- The logic IC 74AC00 must be used as other CMOS logic families are too slow.
- * the .001uF capacitor on pin 1`4 of the 74AC00 to ground must have the shortest leads possible.
- The output can be connected to coax or a small piece of wire as an "antenna".
- Changing the output capacitor to a small value such as 1pF reduces low frequency harmonics.



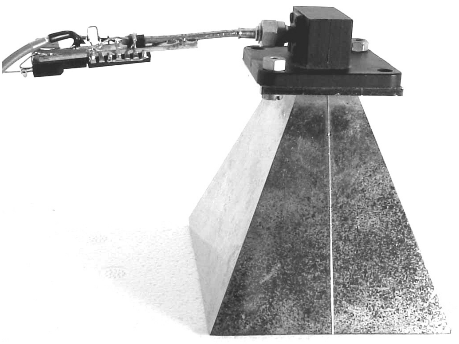
Practical Implementation of the K6IZW Microwave Frequency Marker

Allan, G8LSD

I have made the marker as described Kerry Banke, K6IZW. I had no joy at 10GHz with an 8.5 MHz oscillator but at 24 MHz harmonics are just audible. The good news was when I fed it from an Adret 5104 synthesiser (90 to 120MHz range) I found very pure signals and at very low power. For construction, I used a strip of fibre-glass PCB as a ground plane between the pins. The components were SMDs soldered directly to the pins. The cross connections were made outside and over the top of the pins. For output, I tried a short wire, and this was successful. Then I used a short length of thin semi-rigid soldered to the ground plane, the centre conductor being connected to the output pin by a 2pF SMD capacitor. The output now goes to a transition and a small horn. If the Adret is accurate, my offset on 10GHz is about 5kHz. The short spikes from the 74AC00 are clearly visible on a scope.



Above: the pcb marker Assembly



Right: The marker mounted on small 10GHz horn for test purposes

Probably the Simplest GPS Disciplined Oscillator Possible!

Andy Talbot, G4JNT

GPSDOs

Using the GPS Satellite system offers the advantage of very accurate timing and by extension, frequency control. The long term error is to all intents and purposes zero, with time and frequency accuracy being comparable to the international standard.

The traditional route is to use a relatively low cost GPS receiver module which outputs a 1 Pulse per second signal (1 PPS) aligned to UTC.

Basic GPS modules such as the Garmin GPS25 and Motorola Oncore have been around for some years and are available at low cost. It is possible to phase lock a divided down crystal oscillator to this 1 PPS signal and transfer its long term stability to, say, a 10MHz reference which is subsequently used for deriving any LO and beacon frequencies. The subsequent PLL system is usually described as a GPS Disciplined Oscillator rather than locked, since it is not, strictly speaking, actually 'locked' to the GPS system at all; just controlled by it via the 1 PPS generated by software in the receiver module.

The first GPSDO to appear in the amateur press was by Brooks Sheera W5OJM, described in QST July 1998. A lower stability, simpler version, suitable only for low data rate signalling on the LF bands, was published by myself in Radio Communication October 2002. Both of these, in different ways, demonstrates the problem with using the 1 PPS signal. On all these receiver modules the 1 PPS signal can have up to 1us variation from pulse to pulse, and this varies randomly. Later modules reduce to a few 100ns, but it is still there. Consequently, for a frequency standard with a short term stability measured over a few tens of seconds, this 1 PPS jitter has to be averaged out over many hundreds or thousands of seconds - so giving very long lock up times and loop tracking constant. Now, as the PLL has a time constant of many tens of minutes or hours, the voltage controlled crystal oscillator has to be stable over this loop time constant - particularly if it is to be multiplied up to GHz where a short term wander of a few Hz (parts in 10^-10) is noticeable. So a good quality oscillator has to be used here - typically an ovened high spec standard in its own right.

This was the approach taken by W5OJM with a microcontroller based digital PLL and loop time constant of hours. My design went the other way, and accepted a poor short term stability for LF use only, where the phase wander over a few seconds was inherently averaged by the LF signalling interval. Many manufacturers now offer off-the-shelf GPSDO modules with varying specifications between these , ranging in price from a few hundred pounds, to thousands.

The Jupiter-T Solution

Which brings me onto a new GPS module that makes a homebrew solution very much easier. The Jupiter-T module made by Navman (originally Conexant) has an output at 10kHz 'locked' to GPS time. Initially I was sceptical, thinking it probably only consisted of 10000 pulses per second, which could have been no better than the 1 PPS signal itself in the short term. However, after making extensive measurements, I came to the conclusion that it really was quite a respectable signal. In particular, I could not detect any discrete sidebands at sub Hz frequencies.

This suggested a simple GPSDO solution. By taking a simple, low cost 5MHz voltage controlled TCXO (VTCXO) module and dividing down to 10kHz, this can be phase locked to the output from the Jupiter in an analogue PLL with a time constant of a few tens of seconds.

The circuit diagram **Figure1** shows how simple this can be. Obviously, without the ability to be able to tell if the GPS receiver has locked up by reading the NMEA or binary data it sends from its communication port, there is no way of knowing if the system is functioning properly.

The Jupiter module does output its 1 PPS signal and a nominal 10kHz immediately after switch on, but the timing of these is way off and the initial case of no GPS lock can be inferred from the large frequency error. In fact this is so large that the PLL is out of lock range and the resulting frequency is sweeping so wildly that it is obvious. When the module does lock up to GPS after a few minutes, the frequency and phase of the 1 PPS jump immediately and abruptly to their correct values, with the PLL taking a short time after this to stabilise. Although not shown in the diagram, an LED connected to the phase pulse output of the 4046 chip will slowly change brightness over a few seconds during the GPS lockup, and then much more slowly as the PLL locks, eventually settling to a stable half-maximum brightness

The R/C values forming the PLL filter are optimised to my particular VTCXO which has a sensitivity of 125 Hz/V at 5MHz, and a required tuning voltage in the 0 to 1.5 Volt region. As the bandwidth and tracking performance of the PLL depends on this filter, it is worth spending a bit of time optimising the values in this area.

Some Test Results

The 200th harmonic of the 5MHz output, at 1GHz, was monitored on a UHF communications receiver in CW mode and the output at 1kHz monitored with Argo to show short to medium term frequency shifts. All the local oscillators in the receiver were locked to a Rubidium frequency standard that has been calibrated to an accuracy of a few parts in 10^{-11}. At 1GHz, a frequency shift of 1Hz corresponds to 10^{-9} frequency error.

Figure 2 shows the plot after the system has locked up and been running for about 30 minutes.

It can be seen that the frequency is maintained usually within a couple of Hz, and randomly wanders over a mean period of something like 20 - 30 seconds - this being a function of the PLL bandwidth. The breadboard which produced these results was lying open on the bench, and susceptible to perturbations when I touched it - it is quite possible this trace would be cleaner still if the unit was packaged in a screened metal box.

Figure 3 shows the effect of disconnecting and then reconnecting the GPS antenna. Presumably, the quite fast variation during the period of no GPS signal is the GPS receiver going through its search routine to find the satellites. The faint line that remains fixed at 1kHz exactly is caused by leakage from the Rubidium source controlling the communications Rx.

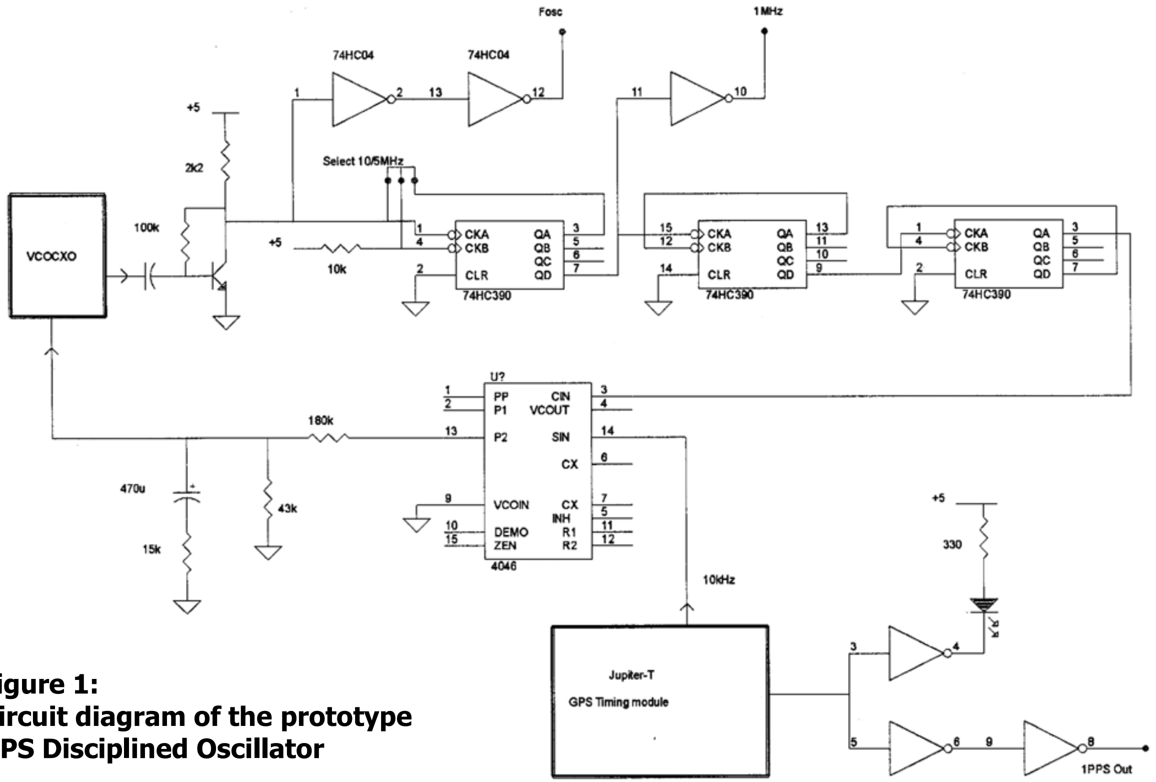
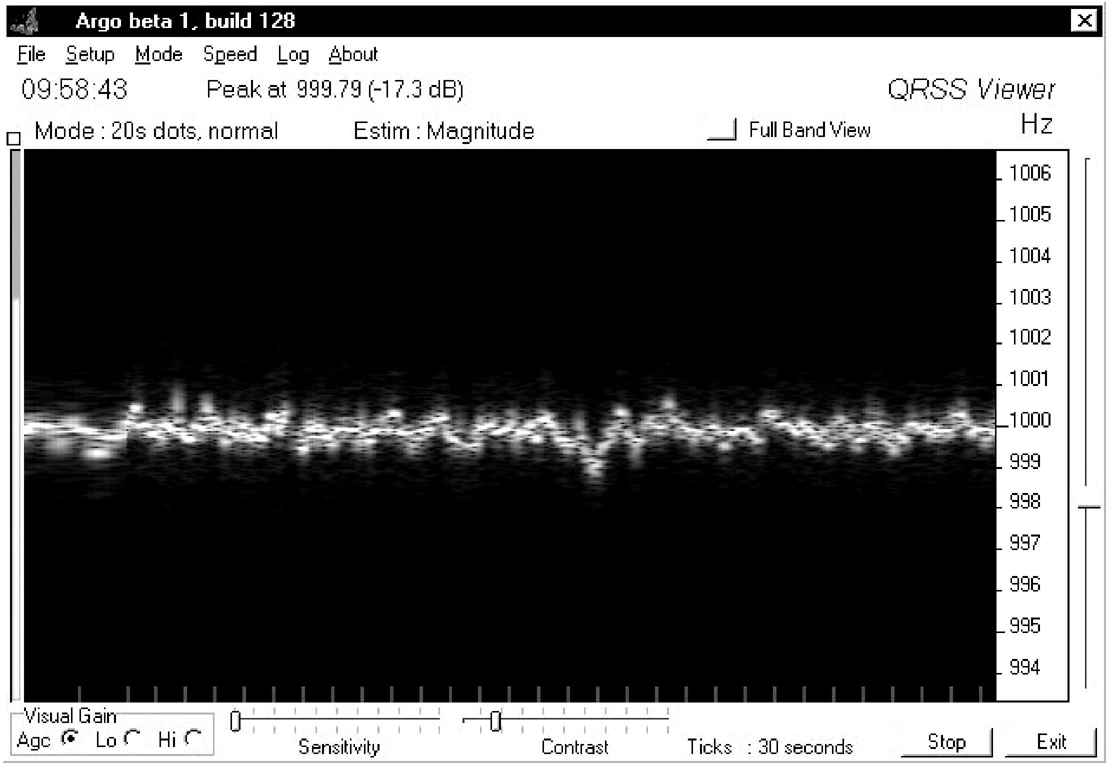


Figure 1:
Circuit diagram of the prototype
GPS Disciplined Oscillator

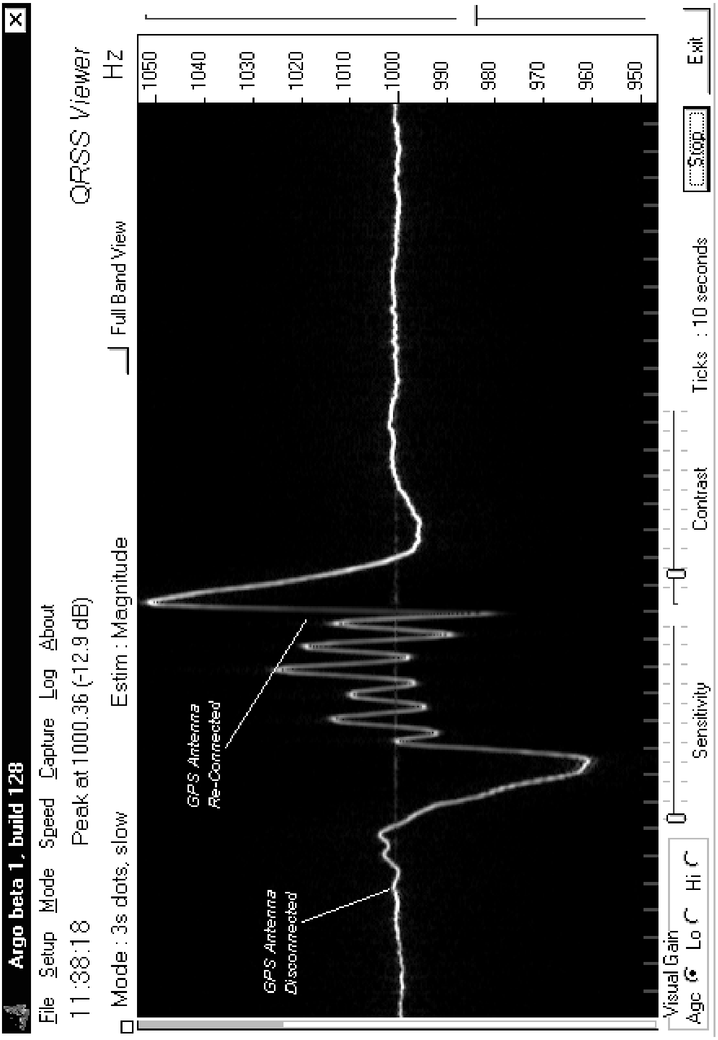


09:58:43 Peak at 999.79 (-17.3 dB)

Mode : 20s dots, normal Estim : Magnitude

Ticks : 30 seconds

Figure 2 Frequency tracking performance when locked, VCXO output multiplied to 1GHz



11:38:18 Peak at 1000.36 (-12.9 dB)

Mode : 3s dots, slow Estim : Magnitude

Agc ● Lo ○ Hi ○

Ticks : 10 seconds

Figure 3 Frequency tracking with loss and re-acquisition of GPS signal—multiplied to 1GHz.

Jupiter-T

This design is based around one specific GPS receiver - others with a 10kHz output may be available, but I don’t know of any. This is the Jupiter T GPS Timing Board.

Off Air Frequency Standard

David Johnstone, GM4EVS

Design notes

This design is based on the G4JNT Mk2 MSF Standard published in the RSGB Microwave Newsletter in April 1996. Changes have been made to the design of the MSF receiver and the 10MHz master oscillator.

MSF Receiver

Good receiver performance is one of the key components of a reliable off-air standard. This is especially true for users located 200 miles or more from the transmitter at Rugby.

Some have used directional frame aerials for MSF reception. Although cumbersome, these can provide an excellent signal. I was fortunate to have an 8in long x 0.5in diameter ferrite rod available.

A winding of enamelled copper wire, 6.5in in length, gave an inductance of some 10mH. This was resonated at 60kHz with a combination of fixed and variable capacitors. This included an allowance of some 180pF for the 6 foot length of RG174 coax used connect the rod to the receiver.

A 12 inch length of 7/8 in OD plastic tubing was used to house the rod, together with a 250pF compression trimmer. The cork from a wine bottle was cut in half to provide a ‘plug’ for each end if the 12 in tube. One of the plugs was drilled to accommodate the RG174 cable.

The receiver itself was built on double-sided PCB, with the top-side acting as a ground-plane. This PCB was housed in a 4.5 x 2.5 x 1in aluminium die cast box. This receiver module was mounted within the main case used for the off-air standard. Good shielding ensures that the receiver amplifies only the off-air 60kHz signal, and not the nominal 60kHz signal generated by division of the 10MHz master oscillator. The receiver circuit followed the ideas proposed by G3LYP to introduce additional LC tuned circuits to improve selectivity. Indeed, those of you with the last 32 years worth of RadCom in the loft, will know that the April 1970 issue provided the design for a Droitwich-locked (then 200kHz) standard by GM3TFY. The receiver used 4 cascaded NPN bipolar LC tuned amplifiers and employed an external antenna.

In addition to the LC filtering, the main FET RF amplifier was changed from an MPF102 to a J310.

Biasing of the FET amp was adjusted to give a drain current of some 8mA. The LC tuned circuits were under-coupled to compensate for the greater gain of the J310. The net result was about the same gain but better selectivity and IMD performance.

During the design, a series LC circuit was included to allow notching-out of a chosen unwanted LW signal e.g., Radio 4 on 198kHz. Use of this was found to be unnecessary in practice.

The biasing of the bipolar amplifier was changed to improve its temperature stability. The emitter voltage was made large relative to the nominal 0.6 volts V_{BE} of the transistor. So when the V_{BE} varies with temperature, the biasing of the transistor remains largely unaltered. Also, some emitter degeneration was used to control the gain of the stage.

The emitter follower design added a 220R series resistor in the base feed to pre-empt any tendency towards oscillation.

The aggregate performance of the

revised receiver was confirmed using the ARRL Radio Designer CAD package. In practice, at a distance of 300 miles, a clean 0.8v p-p output was obtained from the receiver.

With this, the standard remains locked 24 hours per day despite propagation changes at dawn and dusk. As yet, the 10MHz output from the standard has not been multiplied up to check the jitter at 10GHz.

10 MHz Oscillator

Some pulling of the 10MHz oscillator was reported by G4JNT when using the 10MHz output from the Mk2 standard. The oscillator design was revisited. Inspection of the master oscillator circuits for the TS930, TS940, TS850, TS570 and FT1000MP showed them to be all nominally the same. A Colpitts design was used employing a bipolar device having a high gain-bandwidth product. This choice ensures a low base-collector capacitance (C_{BC}) minimising the effect of temperature-driven C_{BC} changes on the frequency of oscillation.

A Colpitts oscillator was built using a 2N5179. This was lightly coupled to a bipolar emitter follower using a 2N2222. You may be inclined to choose an FET follower because of its high input impedance. But overall, the bipolar emitter follower is better-behaved. The prototype oscillator was breadboarded using a small piece of single-sided PCB, face-up, cut into small islands with a hacksaw. The prototype worked so well that it was used for production! The oscillator unit was housed in its own die cast box within the main unit.

While free-running on the bench, the 10MHz oscillator moved about 3Hz in the first 30 minutes from a cold-start. Thereafter, a variation of less that 1Hz per hour was observed during the day with the shack temperature remaining around 64-66F. The measurements were made using a Racal 9916 frequency counter with an OCXO reference that had been on for several weeks.

As the oscillator was operated at low power levels, the 10MHz output from the emitter follower was only 0.4 volts p-p. This level must be raised to 5 volts p-p to drive the 74HC390-based divider chain. This can be achieved cheaply and effectively by wiring a CD4046 as an amplifier – a technique used by Brooks Shera in his GPS-locked frequency standard (QST July 1998).

All sections of the 74HC390 divider were used to give 10MHz, 1MHz and 100kHz outputs. Thereafter the G4JNT Mk2 design was followed. A 3MHz signal was filtered out from the 1 MHz square wave, and divided by 50 to get a nominal 60kHz signal for phase comparison with the off-air 60kHz signal. A CD4046 phase comparator was used followed by an LF353 as a level shifter, to provide the drive to the varicap diode in the 10MHz oscillator. In all, four separate PCBs were used within the unit – MSF RX, 10MHz oscillator, 10MHz divider and phase comparator / level shift. Apart from the MSF RX, the remaining 3 boards were each fitted with separate 78L05 voltage regulators and extensive decoupling capacitors. An external 13.8 volt supply is used to power the unit.

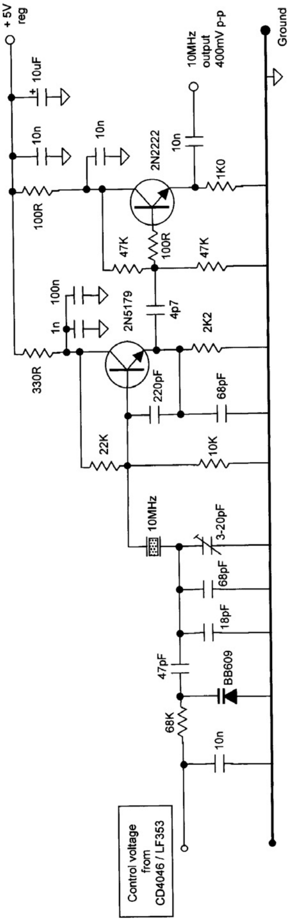
The unit has performed well and credit must go to Andy, G4JNT, for the robustness of his original design. If you won’t always have access to a high-accuracy TV signal, or shy away from the complexity of a digital GPS-based solution, then an MSF-locked standard may well fit your requirements.

Finally, I have some people to thank. These include Dave, G4AON, for his help and inspiration.

We started our MSF projects together, but he built the G4JNT Mk2 design in much less time than I took, and kept reminding me of this!

Also, I am most grateful to Peter, G3PHO, and Mike, G3LYP, for the information and help they provided to make this project a success.

10MHz REFERENCE OSCILLATOR – GM4EVS



Title	10MHz Oscillator – GM4EVS
Author	David N Johnstone
Version	1.1 – as built
Date	30 January 2002

General

Crystal Heaters

Some useful observations by Doug Friend, VK4OE

I recently did some investigations on two Murata crystal heaters that I had available. One type I had bought from Charlie and Petra back in the mid 1980's which I believe were designed for operation at 40 degrees C, and the other type was obtained locally in Australia and intended for operation at 50 degrees C. My friend Rod VK4KZR and I had assumed that temperature was stable, somewhat independent of operating voltage, but some undesirable frequency variations led us to do our investigations. Shown below are the results of our tests. Clearly, the older 40C devices don't vary much with voltage, but 8V would seem to yield the actual specification temperature. But our interest was not much in these 40C heaters because that temperature is often the ambient temperature is Summer here, with internal equipment temperatures expected to be higher. The heaters intended for 50C very obviously should be used only at 5V. At that voltage there is reasonable repeatability between individual devices, which is not the case at higher supply voltages. I can vouch for the very good temperature performance of oscillators using these heaters at 5V. The only drawback is that the crystal should be specified/ ordered for operation at 50C. I have used up the two older 40C heaters in circuits where the crystal has a normal ambient temperature specification. I much prefer the 50C heaters.

MURATA Model	Unit No.	Temp. @ 12.6V	Temp. @ 8V	Temp. @ 5V
BM500N	1	43 ºC	NT	37 ºC
BM500N	2	43 ºC	NT	36 ºC
BG330N	1	58 ºC	55 ºC	51 ºC
BG330N	2	61 ºC	55 ºC	49.5 ºC
BG330N	3	58 ºC	56 ºC	50 ºC
BG330N	4	62 ºC	55 ºC	48 ºC
BG330N	5	63 ºC	55 ºC	49 ºC
BG330N	6	62 ºC	NT	50 ºC
BG330N	7	NT	57 ºC	50 ºC
BG330N	8	NT	58 ºC	50 ºC
BG330N	9	NT	55 ºC	48 ºC

Reducing Warm Up Drift in Local Oscillators

From and idea by David Hall, G8VZT

Murata crystal heaters are commonly used to reduced warm up time and drift in oscillators such as the DDK004. Unfortunately this is not always optimised for minimum drift as the crystal is usually fitted flush to the pc board and the Murata has to heat both crystal case and the surrounding ground plane on the board.

The warm up time can be significantly reduced (by as much as 50%) by simply insulating the base of the crystal can from the pc board. Fit a small PTFE (Teflon) shim or spacer under the crystal, between it and the board. A 1000pF ceramic plate capacitor maybe soldered from the crystal can to the ground plane to effectively ground the can to external RF influences. The ground plane around the crystal can also be removed to some extent in order to reduce the area affected by the Murata heater.

Further improvement can often be made by running the heater off a 9V or 10V regulated supply.

G4DDK004 Oscillator Upgrade

Notes by Kevin Murphy, ZL1UJG

I did some improvements to the last DDK004 oscillator that I built. Listed below are the modifications and the reasons for them. You will need the original circuit diagram and construction notes for reference:

MODIFICATION	REASON
1. Fit a 330nH axial inductor across the crystal (1nF series capacitor Cx must be fitted to prevent DC flowing between TR1 and TR2 emitters)	This gives a stable pulling range due to the xtal capacitance (~7pF) being cancelled
2. Ground TR3, TR4 and TR5 emitters directly. R11, R15, R19, C12, C17, C22, C24 are not fitted. The holes for TRs 3, 4 and 5 were carefully filed with a small round needle file so that the collector and base leads would fold down in little notches while the emitters were soldered directly to the ground plane. The PCB areas where the emitter Cs and Rs were are covered with copper tape and soldered.	This gives a direct path for the emitters and reduces potential problems with emitter bypassing.
3. The 1st striplines in TR4 and TR5 collectors are now directly grounded with copper tape. The transistors' outputs are fed via small chip caps (e.g. 33pF 0805 types). Tr4 has its collector volts fed via a 100nH 1206 size chip inductor. TR5 pcb stripline extended with a small wire. C21, C25 and C29 are not fitted. Where the first tuned striplines in TR4 and TR5 collector circuits are grounded, there are 2 cuts in the 1st striplines. The 1st cut is to isolate the collector from the tuned stripline and the second cut is to isolate the grounded end of the 1st stripline from the Power feed circuit. The 2 pcb cuts can be done in one motion with a scalpel and metal ruler.	This increases the Q of the striplines by grounding them directly.

MODIFICATION	REASON
4. Collector resistors R12, R16 and R20 (associated with TR3, TR4 and TR5) are increased to 150 ohms. The collector's side of each R is fed to the base bias circuits.	This is called a self-regulating, saturated multiplier circuit. (The collector voltage reduces under drive and reduces the base bias). Values other than 150 ohms were not tried. Increased power may be obtained by reducing the resistor values to 120 or even 100 ohms.
5. A 3 terminal regulator (78LO8) is fitted on the pcb with associated tantalum caps of 10uF on the output and 1uF on the input. Note the 4 holes (each 1mm diam) drilled with clearance pads. (use on outboard 7808 1 amp regulator if a crystal heater is fitted and run it off 8 volts).	This reduces the effects of the operating point of TR3, TR4 and Tr5 changing due to supply variations and reduces chirp and other random variations.
6. The small 3 terminal regulator, IC1, already fitted to the pcb for TR1 and TR2 is removed. A new NPN transistor (e.g. BC547 or 2N2222) is fitted (collector to +8V, base to centre hole and emitter towards the Butler oscillator). A 1K resistor from the 8V to the base of the transistor and a 10uF tantalum from base to ground is used. Note: the middle hole must now have a clearance pad so that the base lead doesn't short out.	This reduces regulator noise feeding into TR1 and TR2 circuitry.
7. Fit low cost and more common Philips 1.4 – 5pF trimmers for the 1200MHz multiplier section by reversing the middle trimmer (like an interdigital filter).	Availability and the original trimmers were over coupled, causing the middle trimmer not to tune.
8. An extra 2.2pF NPO ceramic capacitor is fitted on TR3's base and the position of R9 is moved. Two holes need to drilled, with clearance pads.	To improve the multiplier performance (increased collector current)

Notes:

Where tantalum capacitors are mentioned in the original DDK004 documentation and in these modifications, it is very important that they are used as they have superior filtering characteristics. Even if you are not doing these modifications to an existing pcb, it is still recommended that an additional 10uF tantalum is fitted close to the output of IC1 to suppress noise.

The level of crystal harmonic products on the output (e.g. 1200MHz) is better than –45dBc. The board draws ~60mA. MRF901 transistors were used for Tr3, 4 and 5 multipliers and +7dBm (5mW) output was obtained.

R8, 18 ohms, was changed to 100 ohms but it is not a necessary modification.

Overall, the modifications are made to improve the keying characteristics, stability and noise characteristics of the DDK005 oscillator/multiplier. I am upgrading four other units to this standard so that their performance is enhanced for transverter and beacon use.

Since the crystal is sensitive to temperature, it is recommended that a crystal heater is fitted. These are obtainable from Downeast Microwave un the USA or from VK5EME Minikits in Australia. Failing that, some sort of thermal jacket from foam or a polystyrene packing bead can be fitted over the crystal.

For even superior stability, an external crystal oscillator such as that available in kit form from John Hazell, G8ACE, may be fitted and fed into the first stage of the Butler Oscillator, after removing the existing crystal and choke. For matching reasons it is recommended that a J310 FET is fitted instead of TR1 and that R2 is changed to 220 ohms. TR2’s emitter in this case can have the 22pF NPO capacitor replaced with a 1nF type.

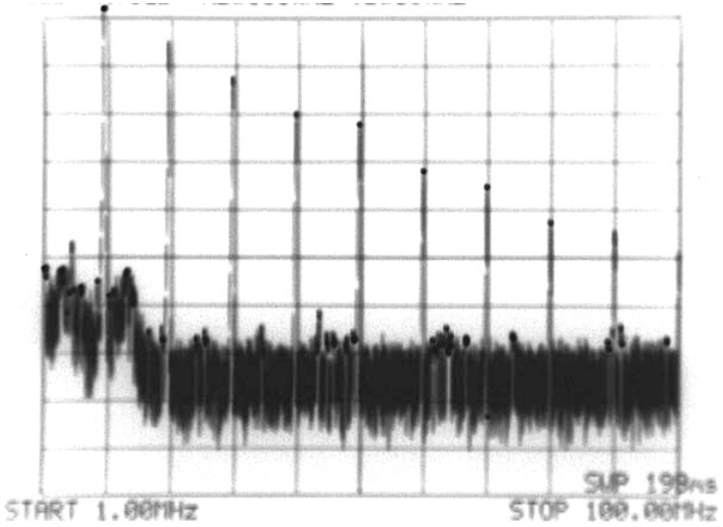
A Distribution Amplifier for the G4JNT PS-DSO

Paul Marsh, M0EYT

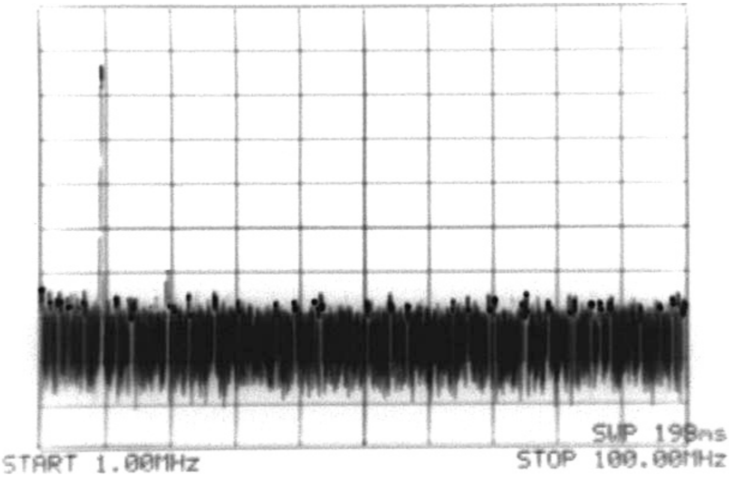
A while back, Scatterpoint published an article by Andy, G4JNT, on making a simple GPS disciplined oscillator based on the cheap Jupiter GPS boards. Since I had a couple of these modules spare, I decided to make one. I used a good quality 10MHz oscillator from John, G8ACE, as the ovened source – then the simple circuit from Andy to lock the oscillator to GPS. The output was locked on 10MHz,and I was very pleased with the results. Then I looked at the 10MHz output on the spectrum analyser.

At this point, I noticed that the 10MHz output and its harmonics extended up over 1GHz – it does make a nice marker generator as is but, since I wanted to lock a receiver to this GPS-DO, I needed to clean the output up a bit. After some hours wasted trying to get good results with multi pole filters I'd bread boarded, I needed a new solution. I looked around the garage for inspiration and found an old box of 10Mbps Ethernet LAN cards that I'd kept. Whilst looking at the isolated BNC sockets on them, I noticed the transceiver modules, so decided to "Google" them and sure enough, one module was a multi-pole filter for the RX and TX paths. After hooking one from the board, I checked it on the spec-an with the results below:

10MHz output before filtering



10MHz output after filtering



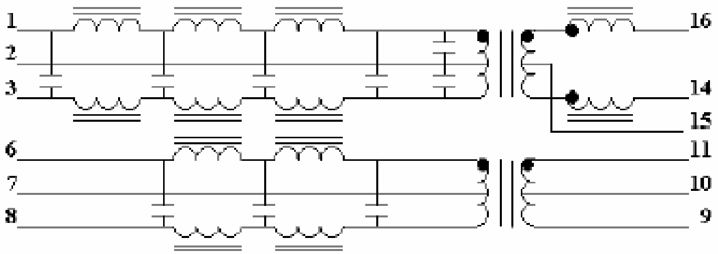
The output from the JNT GPS-DO is first buffered by a 74HC04 Hex Inverter, 2 gates in series, then the second gate driving the remaining 4 gates in parallel.

The 10MHz output from the 74HC04 of course contains harmonics. Each output is fed via a 27 ohm resistor, then into the filter network. The filters are marked 'YCL 20F001N' Each YCL 20F001N package contains one 7 pole TX filter and one 5 RX filter – you can either use the two filters to drive 2 separate outputs, or as I did, wire both filters in series to get further rejection of unwanted harmonics. The pin outs for the filter block are on the schematic diagram below:

The other good thing about these filter blocks are the isolated outputs – if you use the isolated BNC socket commonly found on the old LAN cards, you can get rid of problems caused by earth loops too.

I found once I had built the buffer/filter circuit, the outputs resembled a clean sine-wave on the oscilloscope. I'm sure the output from my old Wavetek synthesiser is cleaner now too. I did make a PCB for this buffer/filter; it is for surface mount DIL packages.

Schematics:



Chapter 2

Frequency Multipliers

A Universal Multiplier for Microwaves

F6BVA

Translated by G3PHO from the French microwave newsletter "Hyper" July 2000 issue

Building equipment for 76GHz has made us review our ideas on local oscillators. In effect, until then, we had been very happy to use the DB6NT kits, which gave us every satisfaction but then, the helical and printed circuit filters are very nice and very effective. The other side of the coin, however, is that that type of circuitry is relatively narrow in bandwidth and does not lend itself to retuning to other frequencies other than those for which it was designed. The specification for my oscillator is quite straightforward:

- Output frequency adjustable from 9.5GHz to 12.8GHz
- • Output power +10dBm (enough for our requirements) with clean spectral characteristics as far as possible (the first spurious is -25dBm

The Local oscillator is made in two separate modules:
The first is a 1250MHz generator, a modified version of one published by F6DER in Hyper some years ago. The frequency s maintained at the correct frequency by using an ovened crystal at around 50 degrees C. Five to ten milliwatts from this stage are sufficient to drive the multiplier.

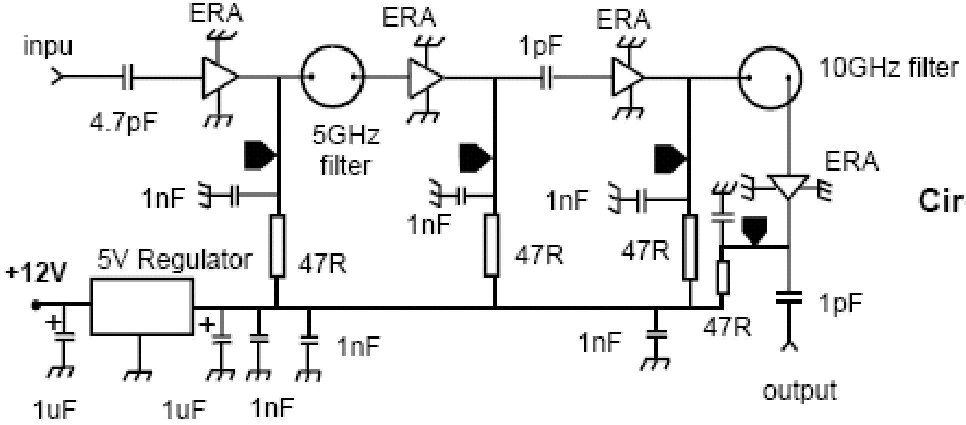
The multiplier forms the second module. An ERA3 is used as a x4 multiplier from 1.250GHz, and produces 5GHz after the first filter. A second ERA3 amplifies the 5GHz and drives an ERA1 which doubles to 10GHz, which in turn is filtered and then amplified by another ERA1. Several examples of the multiplier have been built using the filter dimensions provided in this article and used to produce output between 9.5GHz and 12.6GHz without further modification, other than tuning the filters.

The circuit was etched on double sided 0.8mm Teflon board (Er = 2.35). The filter cavities are soldered to the ground plane side. Prior to this, the copper around the holes for the filter probes is cleared with a 3.5mm drill bit, as are those for the input and output pins of the 5V voltage regulator.

The ground pins of each ERA device are passed through very thin slots (carefully cut with a sharp scalpel) from the strip line side to the ground plane side of the board and soldered on both sides. (editor's note: an alternative method to this would be to use flat head veropins (as in G3WDG and DB6NT modules) pushed through holes drilled from one side of the board to the other. This would then allow minimum ERA ground lead length.)

The filters are made from standard copper pipe fittings ('stop ends' … editor) found widely in DIY stores. The filter pipe caps are then drilled and tapped to take M2.5 tuning screws and locknuts.

The 5GHz filter is 25mm o.d. and



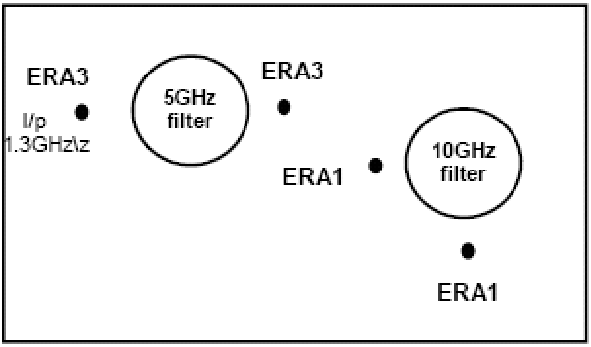
Circuit diagram

17 mm internal height (assuming 1mm wall thickness for the fitting …. editor). The probes for this filter extend 3mm into the cavity, i.e. above the ground plane surface). The 10GHz filter is 20mm o.d. and 10mm inside height. The 10GHz probes extend
2mm into the cavity. (editor's note: to accurately cut probes to length it is suggested they be first cut slightly long and checked with a micrometer. Don't forget to add 0.8mm to their length to account for the thickness of the pcb they pass through.)

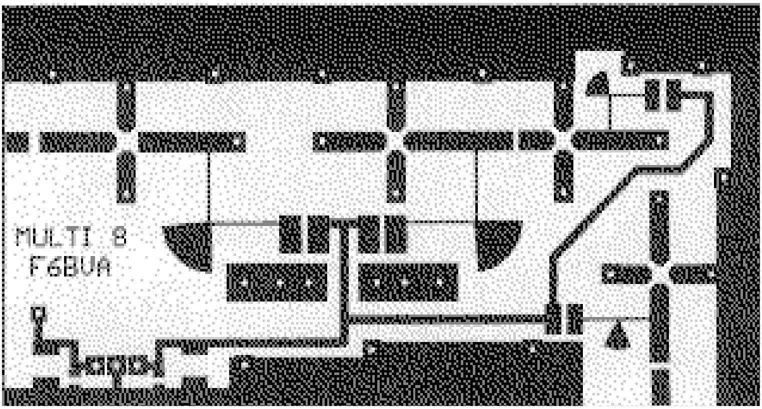
The printed circuit board is mounted in a standard tinplate box measuring 55mm wide x 110mm length x 30mm depth. The ground plane is carefully soldered all the way round at its interface with the sides of the box, 8mm above the bottom of the box.

The construction presented no particular difficulties … just don't forget to make the through connections for the ERA devices. Pre-adjustment of the filters can carried out with the aid of a signal generator if you do not have access to a spectrum analyser. After this adjustment you should see an output of around 5mW. Depending on the chosen output frequency, it may be necessary to "snowflake" the lines with two or three stubs to obtain the +10dBm required. Lossy rubber or foam may be needed in the lid of the box if any instability is found to be present.

This Local Oscillator/multiplier has been fitted into equipment for 47, 76 and 145GHz with entirely satisfactory results.



Physical layout



PCB layout

F6BVA Multi 8 'New Look' Frequency Multiplier

From a translation by John Jaminet, W3HMS

Editor's comments:
The original F6BVA multiplier is described in the previous article. Michel, F6BVA, has now produced this most interesting version, based on cheap and easily obtainable double-sided epoxy PC material. Surprisingly it appears to work as well as, if not better than, the Duroid version!

Introduction
Faced with the difficulties of getting Duroid PCB material, I have redesigned the tracks for double sided epoxy board 0.8 mm thick. Surprisingly ... (it's good that the quality of the substrate is not there for nothing!) … this new design is much more stable than version # 1.

Measured results:
F input 1.0-1.6 GHz at 10dBm
F out from 8.0-13 GHz = 10dBm, minimum and without tuning stubs.
F in 2.0-2.3 GHz . P in = 0 to +3dBm for P out max.

To obtain the total tuning range, the size of the pipe cap resonators have been modified from the previous design.

5GHz Resonator: Interior diameter is 22 mm and the internal height is 7 mm. Probes are 3 mm long.

10GHz Resonator: Internal diameter is 18 mm and internal height is 5 mm.. Probes are 2.2 mm long.

Procedure for Mounting.

- Cut the PCB to fit your tin plate box dimensions. Smooth one cut side edge.
- Drill 0.6mm diameter probe holes through to the ground plane.
- Using a 3mm drill, held in the fingers, cut away the copper on the ground plane side of the board around the holes for the in and out pins of the regulator.
- Drill a 2.2 mm hole for the ERA placement in a way that lets the pins make a good ground contact without being bent.
- From the ground plane side, determine the axis for the pipe cap probes, then use a pair of compasses to mark the mounting position for the pipe caps.
- Solder in the probes and then cut them to 2.2 mm for 10GHz and 3 mm for 5GHz.
- After the pipe caps have been cut to the right height, drill and tap them for the tuning screw, preheat, position them as precisely as possible at the mounting location on the PCB and then solder.
- Solder the PCB into its box. (assume open at bottom for testing)
- With the aid of a fine wire (0.3 to 0.5mm diam.), pull each wire going through the 0.6 mm holes from the component side of the PCB to the ground plane side via the 0.6 mm holes and solder it on each side. Solder the ground wires on the PCB side when soldering the rest of the components.
- Solder all components, MMICs last of all.

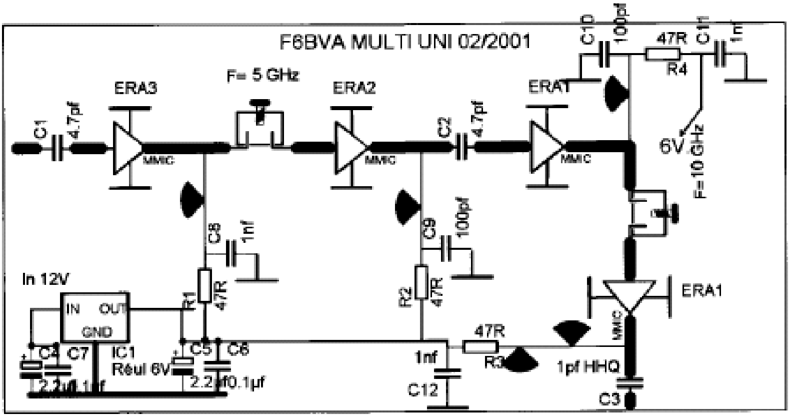
Remember:

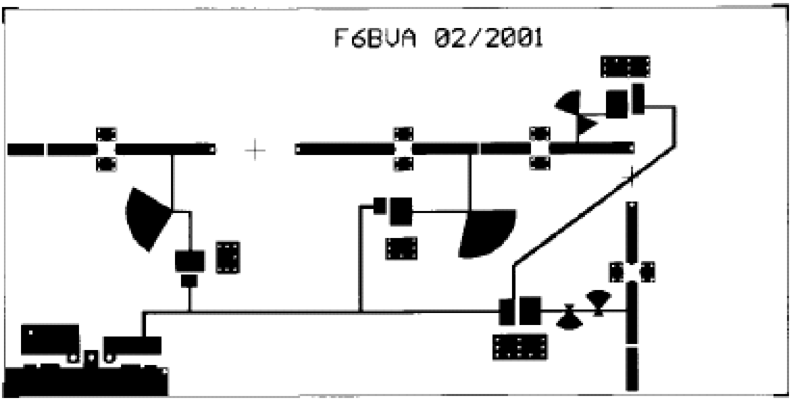
- Pipe caps from 4 – 6.5 GHz =

internal diameter of 22 mm and height of 7 mm with internal probe height of 3 mm.

- Pipe caps from 8 -13 GHz = internal diameter of 18 mm and inside height of 5 mm with probes 2.2 mm.
- Input capacitors at pins 1 and 2 = 4.7 pF. Output capacitors are 1 pF high quality capacitors for use at 10 GHz (e.g.: ATC chip caps ...editor)
- Regulated voltage = 6 VDC and current drain = 200 ma.
- The MMICs as used on the last prototype are, in order from input to output, ERA3, ERA2, ERA1 and ERA1.
- Don't exceed 10 dBm input. Reduce same to +3dBm if you go on 2.xxx GHz.
- Be particularly careful with the MMICs' grounding.

The PC board measures 109.5mm long and 55mm wide when full size.
If using this page to make a board please check these dimensions and adjust on your copier or scanner to suit.





The Qualcomm 2GHz to 10 or 12GHz Multiplier

Chuck Houghton, WB6IGP
San Diego Microwave Group

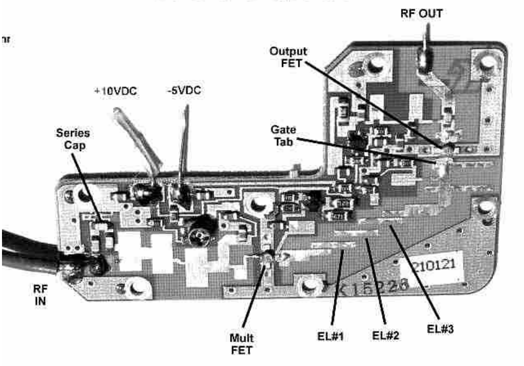


Figure 1. Unmodified Multiplier Board

The Qualcomm Multiplier is used to multiply synthesizer outputs in the 2GHz frequency range by X4 or X5 to the 10-13GHz frequency range. As-is, the multiplier's pc board was designed to multiply 2.620GHz times 5 to 13.1GHz, with a 2GHz drive level of +10dBm to provide +7dBm of output power at 13.1 GHz.

The pc board is quite small measuring 1 5/8 inch by 2 3/4 inches. The board is populated with two stages of MGF1302 FETs and two dc control transistors, one for each FET. The first stage multiplies to 13.1GHz, driving a stripline filter to the second stage output amplifier. The DC power requirements are minus 5 volts bias and plus 10 volts dc. Figure 1 shows the unmodified pc board and indicates the areas involved in this modification.

This board is a natural to drop into a 2 by 3 inch housing for frequency multiplication schemes in the 10 to 13GHz frequency ranges. Our earlier Qualcomm output filter modification schemes for 10GHz used a 1/2 inch pipe-cap filter in place of the original stripline output filter. That board was nearly 3 times larger than this newer smaller multiplier board. This means that you can have a 2.x GHz multiplier to 10 or 12GHz in a small aluminium box of 2 by 3 inches by 1/2 to 3/4 inch high. The filter approach for the new small board is to use standard stripline extensions made of copper strips the same size as the original stripline filter elements. Extending the element lengths lowers the resonant frequency from 13.1GHz to the new multiplier output frequency range.

The Modification

The new dimensions in this modification are specifically to multiply a 2.556GHz input X4 to 10.224 GHz as an LO for 10.368GHz operation. You will add a 1.5 pF chip cap to the existing input filter.

This improves the drive to the multiplier stage in the vicinity of 2.5GHz. Modifications to the multiplier circuit output filter consist of extending three stripline filter element lines by strips of copper soldered to the original stripline elements. This retunes the filter for best output at the 4th harmonic for a 10.2-10.4GHz LO. The gate tab improves the output by about 1 dB. Typical output power can be +4 to +8 dBm. This of course depends on drive and re-tune success. If you need to have a different output frequency range than the 10.224-10.368GHz described above, then the output filter element lengths and gate tab dimensions will need to be experimentally redetermined.

The multiplier pc board has a notch in one corner about .75 by 1.5 inches in size.

Although I didn't take advantage of this space in the modified unit in this article, the notch allows room for a small internal power converter in the same enclosure. The power converter could provide the 5V negative bias of 1 ma or so, and regulated +10 V, allowing the multiplier to be self-contained operating from a +12 volt dc main supply.

Again, as with all modified Qualcomm material, DO NOT CONTACT QUALCOMM for information on these devices. Please contact Chuck Houghton, WB6IGP, clhough@pacbell.net, or Kerry Banke, N6IZW, kbanke@aol.com for any information or availability. More details are on the Web at :

http://www.ham-radio.com/sbms/sd/multx4.htm

Chapter 3

Antennas

Antenna Measuring Notes

Kent Britain WA5VJB

Since 1987 I have set up my portable antenna range at 14 Conferences measuring well over 1000 antennas, mainly in the .9 to 24 GHz range. G4DDK has asked me to list some of my observations.

The Feed is not at the focus of the dish:

First off, I have NEVER been able to calculate the focal point of my dish, mount the feed, and have the antenna optimised. NEVER! It always seems I have to move the feed in towards the dish a bit to tweak things up.

But out of the antenna range things are far worse. About half of the dishes have the feed off by as much as 50% in distance!

A chap comes up with a 2 ft. dish and about a .35 f/d. The feed is sticking out 3 ft from dish! "But that's where I calculated the focus to be!" is always the answer. I haven't found out what in the d sq/16c equation throws them but we see it all the time. Another problem is the rounded edge on most dishes. They measure the physical diameter of the dish, not the diameter of the actual parabolic surface.

That outer cm or so is not usable and should not be used in the F/d calculations. And I won’t even start on the complications of calculating the actual phase centre of the feed. I have always been able to pick up a dB or two tweaking the focus and 6 dB or so has been the typical improvement at the conferences when the feed is movable and we can optimise it's position. Finally come the 25% or so really bad ones. The "dish" was not parabolic, the feed wasn't resonate in the ham band, the focus was miscalculated by 3 feet (on a 2 ft dish!), using a grid dish on 3cm (and 3cm spacing on the wires), and so on. As I said, typically 25% or so of the dish antennas tested at these conferences are just air cooled dummy loads.

Most of these really bad ones are usually the prime focus dishes. With the offset dishes, you usually have a pretty good idea where the feed was for a starting point if you had saved all the parts. But the ham feeds will usually have a different phase centre, so we're back to seeing several dB of improvement moving the feed around a few millimetres. The antenna range is also a good spot to figure out where the antenna is pointing. Build up some kind of mechanical sight, or mount a telescopic sight on the edge of the offset dish. (Top edge works well too) Peak up the signal, and sight in on the source. An optical sight can be very useful when portable.

On Yagi and Quad loop type antenna, the builder has often used a different diameter boom, or different size elements, or even replace a round boom with a square boom. In each case there was no attempt to use corrections factors. (50/50 chance they would have gone in the wrong direction anyway!) Again a large percentage of these antenna are not close to what the owners are expecting in the way of performance.

If you tested all the microwave antennas in England, I am sure very few of the antennas would have the dBs you have been optimistically using in your range calculations.

Interestingly both the UK and ZL groups left the Microwave Update antenna range saying, "This antenna measuring is not all that hard" and are both looking at setting up antenna ranges back home.

Next month I will go more into the details of measuring antennas and some of the pitfalls that have given me extended headaches. It really is not all that difficult to set up a practical UHF + antenna range. Last month we discussed the value of an antenna test range, now we will explain how it is not necessarily complicated to set up an antenna range.

Equipment:

RF Source
Source Antenna
Reference Antenna
Detector
Open Space

RF Source 1000Hz vs. CW:

Many of my first antenna range setups was just a CW source, a reference horn, and a power meter.

Hold up the power sensor and horn, measure power, attach the antenna to be measured, and the difference is dB power is the difference in dB gain. This works, but you really have some dynamic range problems. The power sensor is not very sensitive, so you have to run a fair amount of power and use a short range. But it does work. Just make sure you have at lease 1 dB more power that the noise floor of the power meter, otherwise you run into (Signal + Noise) / Noise problems. I have been able to make pretty good measurements with 10 to 30 milliwatts sources into 20dB gain antennas on 6cm and 3cm. Horns were measured at about 5 Meters, dish antennas at about 10 Meters from the source.

Generating 1000 Hz:

I have a Wavetek 3001 50MHz synthesised signal generator I haul to the antenna range. Most RF generators already have a 1000Hz AM setting. So up to 50MHz I just set the generator to max output, 100Hz AM and drive a source antenna with it. On 902MHz, I set the generator to 45 MHz and drive two sides of a mixer, this doubles to 902MHz which goes through a filter and into a 20dB gain amp. This gives me about 100 milliwatts to work with. On 1296MHz I set the generator to 432MHz, horribly overdrive a small brick amp, filter the 3rd harmonic, and drive a second brick amp.

Again about 100 milliwatts to work with. On 230 MHz, I set the generator to 384MHz, again horribly overdrive a small brick amp, filter the 6th harmonic, and drive a second brick amp. This gives me about 75 milliwatts to work with, and there is a second similar unit for 2400MHz. For 3456MHz I again use a brick amp (Hand picked this one) driven with 432 MHz and run the 8th harmonic through an interdigital filter. A second brick amp brings this up to 10 milliwatts or so. On 5.7GHz, 10.3GHz, and 24.1GHz I use Gunn sources driving PIN Diodes. Sometimes I use a 555 timer circuit AA5C built up for me, other times I just drag along a function generator and directly drive the PIN Diodes. On 47GHz I have a 23.5GHz Gunn source driving a doubler out of an old HP 940A . I modulate a PIN diode on the 23.5GHz source.

So there are a lot of ways to generate a 1000Hz modulated RF source. Of course if I had a Signal Generator actually on these frequencies, or even a sweeper with an external modulation input, I would use it.

Source Antenna

You will need an antenna at the source end. It's nice if the antenna has a fair amount of gain and over the years I have used everything from Coffee Can horns, to 2 ft dish antennas on 3 cm. Over the years I have migrated to multiband antennas at the source, just to speed set-up and less stuff to haul about.

Ridged horns work well, some the multiband dish feeds work well too. While it is nice for the source antenna to have gain, it is not necessary and I am a firm believer is using what works. More on the type of antenna to use under Open Area.

Detectors/terminations

At the receiver end we need a simple diode detector to demodulate the 1000Hz AM signal. I normally use a standard Type N Input diode detector. Now, most of these detectors in the US do not contain any kind of terminations. So they don't look like a 50 ohm load, but rather have a very complex input impedance. Just put a 6dB pad on the input of the detector. 10dB works better, 20dB has too much loss, and the input is pretty much 50 ohm. For the higher bands, the simple diode detector mounted in WG works well.

Receiver:

If you are using CW and a Power Meter, then this is your receiver. If you are using 1000Hz, then you need an HP-415, HP416, or the Marconi Type 6593A. Many other companies also make these 1000 Hz "SWR Indicators". These are simply an audio meter tuned to 1000 Hz with a high accuracy meter scale. I have also used General Microwave and NARDA versions of the HP415 and they work just as well.

Note: The HP415E needs 9 Volts to run, 16 volts to run in the expanded scale mode. It only pulls about 5 ma, so I will wire in two 9volt dry cell batteries and can run it for hours and hours on the antenna range.

Reference Antenna:

The most important part of the antenna range is an antenna you know the gain of.

The hand of God, or someone with a crayon, has written the gain of the antenna on a calibration sticker, or Post-it-Note. We measure how much signal the Reference antenna collected, and if the antenna being tested collects 3.2dB more signal, then it has 3.2dB more gain than the Reference. On all bands above 1.3 GHz, I use horn antennas as the reference. In several cases I was fortunate to acquire Std. Gain Horns, and the gain of a horn antenna can be easily calculated with high accuracy.

But like any ham activity, absolute gain numbers are not as important as optimising the antenna. We will spend hours tweaking the output of a power amp, or the NF of a Pre-amp, and we

know the equipment is not calibrated for the frequency we are using but Max Power Out is still Max Power Out wether it is 68 Milliwatts, or 92 Milliwatts, it's all we can squeeze out of that circuit. Same for the antennas. You move the feed around, try several feeds, and so on, even if the range errors are a dB or 2, we have maximised the performance of the antenna and been able to compare the relative performance of different antennas. Again if the real gain of the antenna is 29.6 dBi or 30.4 dBi is all academic, we have squeezed everything we can out of the antenna. (Or thrown it in the rubbish bin!)

Audio out:

The HP-415 has a "Recorder Out" connector on the back. I usually connect a small audio amp with a speaker to this connector. The raw 1000 Hz can be amplified to drive a speaker. Several good reasons to listen to the audio signal. It's nice to peak the antenna on a audio signal, especially when you're holding the dish with one hand, moving the feed with another hand, and holding the detector with the third hand .... you know what I mean. And it is especially good when there might be interference Some time back I was testing an L-Band Helix for possible use on Phase 3-D. The meter was jumping all over the scale, there were several peaks away from the source, an a constant erratic noise floor on the meter. After some time I

connected the audio amp and figured out the problem in nano seconds. Loud video buzz! The long helix was acting like a 1/4 wave whip on TV Channel 4! All further testing included either a 1269 MHz filter or an isolator. I have had similar problems testing Log Periodics that do not have the back ends of the booms terminated. At our Central States VHF Conference we typically have the 50-450 MHz and the 900+ MHz ranges running at the same time. Listening to the tones lets us easily tell when of the English lads were looking at me pretty funny while I was waving a horn antenna all over the parking lot at Microwave Update. (The US guys had seen me do this before.) Up, Down, Left, Right, Back up a bit. The antenna range is not preplanned geometry, I am just looking for an area about 1 meter by 1 meter where the signal level varies less than 1 dB. When I find it, I put some kind of marker on the ground, then tell everyone how high to hold their antennas. Over grass I will usually set the source antenna 2 or 3 meters off the ground. If the source antenna has 20 dB or so of gain, very little of the RF hits the ground at the 1/2 way point. And what does hit the ground is higher than Brewster's angle, so the bounce is attenuated 10 to 15 dB. If I am on a parking lot, I will set the source on the ground and make a ground reflection range. Although there is no hard fast rule here, I just use what ever works best and it's a quick test just setting the source antenna on the ground. If there is a nice consistent signal area, we start measuring antennas!

Measurement technique:

Substitution: Measuring itself if simple and quick. Hold up the reference antenna, set the meter to a convenient sport, attach the sensor to the antenna to be tested, hold it at the same spot, take a reading, calculate the difference. I normally carry some kind of marking pen and write the results on the antenna. We usually have someone else standing around with pen and paper making a more complete record, but the guys seem to like having an "Official Result" right on the antenna rather than trying to remembering it, or waiting a month until someone publishes the results.

Dynamic Range: One pitfall is the dynamic range of the SWR Indicator or power meter. You like to keep less than 10dB difference between the antennas under test. So don't use a dipole as the reference for a 30dB gain antenna measurement. First of all the meters have errors the farther you stretch them.

Second, the capture areas of the two antennas will be quite different. A large number of secondary problems testing antenna with vastly different capture areas. With the 415's or 6593's you want to keep them down in the 30,40 or 50 dB ranges. Higher than 30 usually means the diode is driven out of the square law region, in the 60's the signal will be pretty noisy. These meters will also work with bolometer mounts. Now you could use all the scales with a bolo, but the bolo is less sensitive than a diode mount and you will need more signal. The 6593 can be used to directly compare 2 antennas, but this means you will need to find a larger measurement area, bigger than both antennas, to make your measurements. This is easy enough on 50 MHz - 432 MHz, but much more difficult on the microwave bands. I haven't used a 6593, but going over one in G4DDK's garage, it sure looked like a natural for antenna ranges.

Results: Oh it was fun in the early years deflating egos. "Well, a 5 element Yagi would have 12dB gain, using quad elements adds 2dB, and a corner reflector would have 10dB, so by combing a Yagi, Quad, and corner reflector, my super antenna has 2 dB GAIN!" Yea, sure, here's the detector. (6dBi if he was lucky!).

Over the years the wild claims have died down, and better, more consistent designs are showing up. And we have developed a bit of a tradition of seeing what kind of strange antennas we can show up with and still get good results. And a sprit of experimentation has developed where guys are not afraid to show up with a dish, 8 feeds, and find out which one works best. Typically at the CSVHFS antenna contests we will measure 100 to 125 different combinations of antennas. There have been a few fun ones, I particularly remember KB0HH spending several years trying to optimise a scalar feed. With excellent form, his cowboy boot set the feed over 30 meters down the range!

Circular Polarisation: We usually get a few CP antennas to test. Normally I just measure the gain, rotate the antenna 90 degrees, measure the gain, average the numbers add 3dB, and label the gain dBiC. Ideally, the gain does not change as the test antenna is rotated. If gain only varies 1 dB I'll congratulate the builder, if it varies 3 dB I'll still call it CP, more than 3dB and we'll start looking at ways to fix/repair/improve the antenna. This is especially a problem with some of the "Short" Helix dish feeds that have become popular lately. It is very difficult to properly generate a CP wave in only 2 turns of wire. At an AMSAT Conference we set up the antenna range and only 4 of the 8 Helix antennas had gain along the axis of the antenna! Of the 4 with gain, only 2 were within 3dB of that 1296 rhombic also working as a 144 MHz antenna. (Our 1000 Hz tones are hardly phased locked)

Open Area:

30 Meters is nice, but I have often set up in more confined areas of 10 to 20 Meters. But try to avoid areas near walls that might cause reflections. Given a choice, I set up on grass (easier on my feet) but parking lots can also be used. The whole idea is to find an area where you have a consistent signal about the same size as the capture area of the antenna. i.e. a bit bigger than the biggest antenna you plan to test. Our greatest source of error is have the signal level on one edge of the dish stronger than on the

other edge of the dish. Some circularity. And yes, 1 of the 2 had been brought by James Miller G3RUH. The AMSAT lads have been passing around the idea that Helix antennas are easy to build and fool proof. Test of dozens of Helix antennas says they are WRONG. AMSAT writers perhaps have the worse habit of copying articles. A guy writes an article about a Helix, that is copied from and article, that was copied from and article .... And over the last 5 generations of this design, each writer/builder has substituted materials, slightly changed dimensions, and NEVER tested the antenna.

Log Periodics: LP's can also be difficult to test, especially the ones that do not terminate the back of the booms. The antenna pick up fundamental and harmonic frequencies equally well. They also tend to pick up more local interference. A clean source, and monitoring the 100 Hz audio will usually keep you out of trouble. The unterminated LPs tend to act like a big capacitor and pick up noise from the mains and a lot of other garbage. All my current LP designs terminate the back of the booms, it just cleans up so many problems.

Higher Bands: In my job, we have been doing some radiometry work between 90 and 110GHz. Yes, I modulate the Impatt amp with 1000Hz and do all sorts of tests with the HP 415. It's a system that works well for antenna testing on all bands.

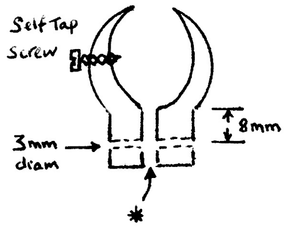
Yagis and Helixes

A ‘Do it yourself’ Yagi for 23cm

Sketches and notes by Peter Blair, G3LTF

Editor’s comments:
This is not a complete constructional article in that the numbers of elements and their dimensions are left to the individual to decide. The beam is based on the DL6WU design (for which there is DOS-based software freely available) and on the article by Rainer, DJ9BV, in Dubus 2/94 and Dubus Technik V page 96. One departure from the printed article is the use of fully insulated elements, mounted underneath the boom rather than on top of it. Anyone with access to a drill press, hacksaw and callipers should be able to make themselves a high performance 23cm Yagi at low cost, if these diagram (and the DJ9BV articles) are carefully followed.

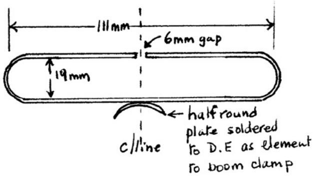
Boom diameter: 19mm O.D round section tubing
Elements: 3.6mm O.D aluminium rod
Driven Element (D.E.): 4.5mm copper rod
Element supports: 15mm plastic pipe clamps (from DIY or plumber)
Each pipe clamp is drilled with a 3mm drill, 8mm from the boom, in the position shown in Fig.1
The clamps are secured to the boom by means of a self-tapping screw.



Each element is a tight, push fit into the 3mm hole and is secured with window sealant injected into the original screw hole in the end of the clamp (see * in fig.1)

The directors should be made 9mm shorter than the dimensions given in Dubus. Lengths should be accurate to +/- 0.3mm. Elements end should be filed flat across the rod end cross section. The Reflector length is 113mm.

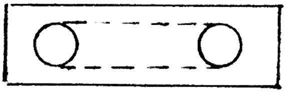
A jig to shape and bend the driven Element (Figure 2) to the correct size is easily fashioned out of two pieces of 19mm OD round bar, mounted on a small piece of aluminium plate as shown in Figure 3.

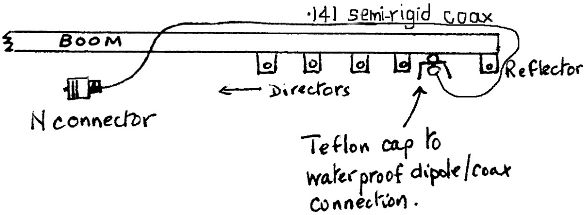


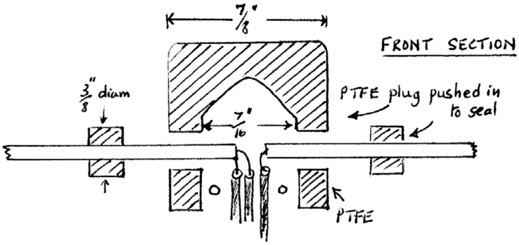
All the elements, including the driven element, are mounted on the underside of the boom. The balun is as the Dubus article fig. 3. The feeder is 0,141” semi-rigid coaxial line, approx, 1 metre long, taken directly from the driven element and back over the reflector as shown in Figure 4 A Teflon cap waterproofs the driven element feed point centre section, as detailed in Figs. 5 and 6.

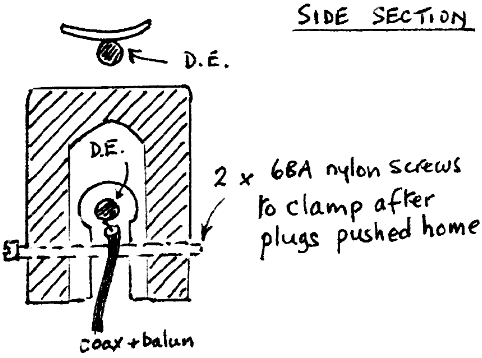
The completed Yagi should be supported about its centre by a "trombone" bracket section as used in UHF TV antennas. This allows the Yagi to stand clear of the supporting mast, which can be, for example, a 2 inch diameter pipe.

MANY THANKS TO RAINER FOR THE FINE BASIC DESIGN!









Heavy Duty Antenna Systems for Portable Microwaves

Peter Day, G3PHO

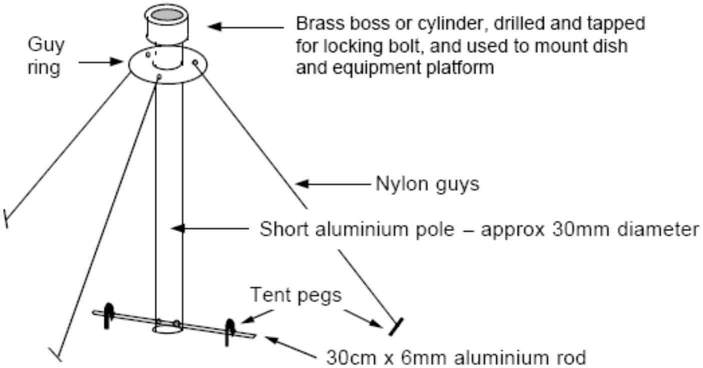
Have you ever had your microwave gear blow over in the wind when out portable? There can't be many portable operators who have not suffered in this way! Since ninety-nine percent of my microwave operating is out-of-doors (and has been for the past 25 years or so) I have gone through many different dish and Yagi support systems. Some have been moderately successful but others have been downright disastrous! The aim of this article is to pass along some of the lessons I have learned over the years, in the hope that they may be of help to others. While these support systems refer to a van rather than a car, there is no reason why a smaller vehicle cannot be used in the manner to be described …. just scale down the masts!

While tripods are still the most widely used dish supports in portable microwave systems, they leave much to be desired, in spite of their obvious advantages of ease of setting up, light weight and compactness. Good tripods are hard to find and are very expensive when purchased new. Few operators own the superb ex-BBC/ITV outside broadcast tripods use by operators such as G0HNW and G3KEU. My National studio tripod, while very sturdy, is not really safe with dish diameters greater than 40cm or so. The present-day tendency is to use offset dishes of 60, 90 or even 100 centimetres diameter. These put a tremendous strain on the average tripod (often not much more than a large camera type). Wind loading, in even light breeze conditions, can be enough to cause the tripod legs to lift off the ground. Unless the tripod is securely guyed, stronger winds can easily tip the whole system over, causing the equipment to crash to the ground and be damaged in the process. I have never really felt confident with tripod-mounted gear and so have tried a variety of other methods over the years. These have involved guyed masts, both short and tall and, more recently, guyless masts supported by the vehicle.

Guyed masts:

For lightweight, backpacking forays into the hills you can't beat a short length (say 1.5 metre or so) of aluminium tubing as a dish support. This can easily be guyed with three thin nylon cords and tent pegs. The base of the tube is drilled through to take a 60cm length of 5 or 6mm diameter aluminium rod which, in turn, is anchored to the ground with a couple of tent pegs. As this prevents the mast from turning, a dish-to–tripod rotating head mount needs to be fashioned to slip over or into the top of the short mast. A locking bolt tightens the antenna mount when the direction has been determined. **Figure 1** shows the essentials of the system. Of course the mast can be as high as you want it. The writer has used a 3 section, 4 metre mast, with a 48cm dish clamped at 1 metre and a small 144MHz Yagi at the top. Such a system was used extensively in the days of 10GHz wideband portable when it was thought essential to get to the highest summits possible! There is still a need for this type of lightweight portable today, with the advent of 47GHz and 76GHz activity. These bands need all the help they can get so "the higher the better" still applies.

FIGURE 1


Brass boss or cylinder, drilled and tapped for locking bolt, and used to mount dish and equipment platform
Guy ring
Nylon guys
Short aluminium pole – approx 30mm diameter
Tent pegs
30cm x 6mm aluminium rod

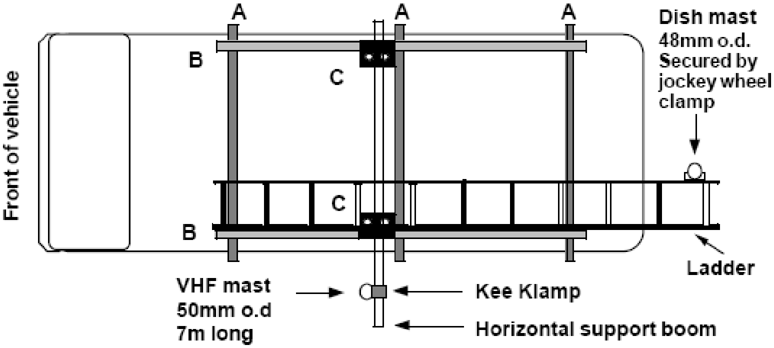
Such a mast weighs very little and can be easily strapped onto the side of a rucksack, along with a small dish. For these "high altitude" operations the 144MHz talkback equipment need be nothing more than an FT290 or IC202. Provided the output power of the microwave station is in the order of milliwatts rather than watts the battery requirements can be met with Yuasa type "dry fit" batteries rather than the large leisure ones used with high power car-based operations.

Vehicle supported masts

Over the past twenty years I have operated from a VW Campervan during the all-day microwave contests. This has the advantage of lots of storage room for equipment, a table and seat operating position and the satisfaction of remaining out of the wind and rain during bad weather. The idea of standing outside all day, by tripod mounted gear has never appealed to me, even though I am quite used to Scottish mountain winters! In February this year I acquired a new vehicle, a VW Transporter. It was a "bare" panel van when I received it from the dealer but I have since floored and carpeted it as well as lining the interior walls with fibreglass wool, plywood and carpet. Side windows were then fitted in the sliding door and the wall opposite. No rear seats have been fitted as this vehicle is for microwaves only! A small operating bench is bolted to the wall beneath the window. I fitted three VW roof bars to the vehicle and added two further longitudinal, square section bars down from the front to the rear roof bars, thus forming a very rigid frame (**Figure 2**). A metal plate, 19cm x 10cm, was then fitted midway along each longitudinal bar and drilled to take U-bolts for a 3 metre long, 50mm diameter, aluminium pipe that runs under the frame from one side of the vehicle to the other. This forms a horizontal support for a 7 metre, 2 section mast that holds the 8 element 2m beam for talkback.

No guys are needed as the mast as it has 3mm thick walls and is approximately 50mm diameter. I cut it into two sections and use a joining sleeve for full height. If conditions are extremely windy I can always use just one section but things have not yet come to that! The vertical mast hinges on its support boom by means of a Kee Klamp, a 90 degree joiner used in scaffolding and barrier structures (**Photo 1**).

Figure 2: showing roof bar structure and mast arrangements



A: Roof bars (spaced approx. 1 metre apart)
B: 20mm square section steel tubes, secured to roof bars with M6 bolts
C: Mounting plates, drilled for 50mm U-bolts and secured with M6 bolts to square section tube

The base of the VHF mast slots over an axle stand (I have two in my garage … they are rarely used for car servicing!) to allow for easy rotation. The mast is locked into position by tightening the hexagonal bolt in the Kee Clamp. The mast in the foreground supports the 144MHz beam. A further Kee Klamp holds a turning handle at a suitable height. The 1.2m microwave dish antenna is not fixed to this mast as it is on a separate mast fixed to the ladder.

The microwave antenna mast is supported at the end of a short, 4 metre ladder section which is clamped to the van roof bars. Here standard ladder clamps are employed. These allow a quick but secure attachment of the ladder to the roof bars. Around 1 metre of ladder projects beyond the back of the vehicle and a 3 metre high mast, consisting of 5mm wall aluminium, 48mm o.d., is attached to the side of the ladder, near the end rung. For this I use a caravan or trailer jockey wheel clamp, obtainable at low cost at any caravan spares supplier. The clamp is bolted to the outside of the ladder so that the ladder can still be used for climbing if need be. **Photo 2** shows the clamp in close-up. Note the sturdy locking handle and general rugged construction. The

Photo 1: VHF mast and support

jockey wheel clamp allows both a secure support for the 3 metre mast and a convenient means of locking the mast in position when dish headings have been set.

The mast stands in a further Kee Klamp fitting which is mounted on a wooden board upon which a 360 degree compass scale was carefully drawn and protected with clear varnish. The board is anchored to the ground by means of tent pegs pushed through holes in each corner (**see Photo 3).**

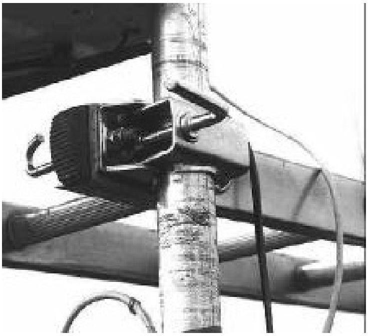


Photo 2: Microwave mast support

A simple direction pointer was made from a strip of aluminium , painted black and clamped to the pole as shown. The 1.2m dish has a mounting ring and this is used to attach it to the top of the support mast such that the dish can rotate through a full 360 degree arc, above the van roof. The 10GHz transverter is carried on a small platform attached to the mast, just below and behind the dish. A short length of WG16 flexiguide transfers the RF to a Chaparral" type feedhorn that came with the dish. Once the mast and dish are elevated into position and clamped to the ladder, the transverter is placed on its platform and secured with a couple of "bungies". A 3-way spirit level (local D.I.Y again!) is then strapped to the mast and everything is trued up. With 5 watts of RF output, a 1.2m dish and a HEMT receiver front-end, I now find that I can work 400km or more on 10GHz with relative ease, if conditions are anything like normal. I have the distinct impression that the dish performs better at the new height of approximately 3 metres above ground. Plans are now afoot to use the same dish on 5.7GHz with my newly constructed DB6NT 6cm transverter.

With a "bombproof" antenna support system and a roomy vehicle, I can now confidently face anything the British weather throws at me (says he, with fingers crossed!). I hope other readers might find some of this information of use and that we might see more dedicated portable microwave operators out and about in the various activity days and contests.

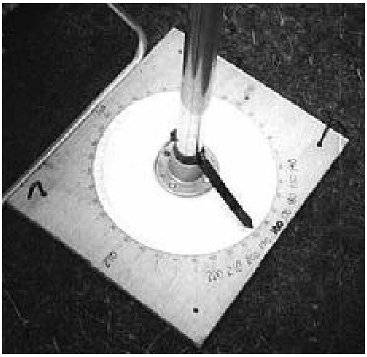


Photo 3: Microwave base board

HARDWARE RESOURCES:

4m ladder section: Local D.I.Y store. Look for light weight aluminium type,

Kee Klamps: Excellent catalogue (Fitting Manual) available from Beeley Fabrications Ltd., Niagara Road, Sheffield S6 1NH (Telephone 0114 234 3244 Fax 0114 2343063)

Roof Bars: VW main dealer.

20mm square section steel tube: Local B&Q.

Dish mast (48mm o.d/5mm wall):

Local scaffold supplier at £2 a foot.

VHF Mast (50mm o.d/3mm wall) and joining sleeve: Local TV antenna supplier at £1.25 a foot.

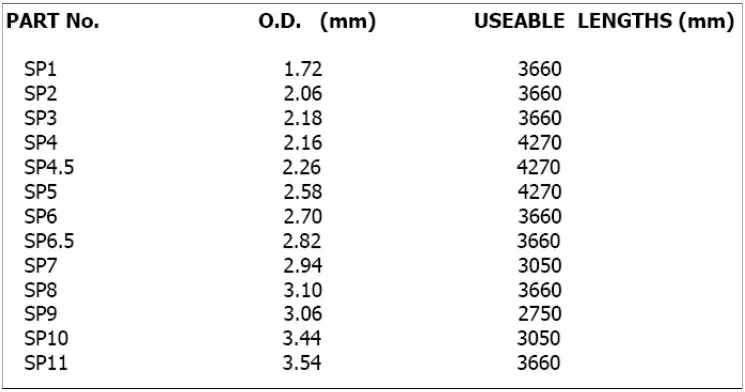
Jockey Wheel clamp: Local caravan suppliers. Approx £3.75 for large size.

Making Helix Antennas

some notes by Des Clift, VK5ZO

Over the last few years I've made quite a lot of helix antennas from 23cm up to 3cm on 2.4GHz, 3.3, 4.5, 5.7, 8 & 10GHz. I solved the problem of obtaining suitable wire by using capillary tubing service packs that are used by refrigeration people. They are made in Adelaide and are available in 12 or 13 sizes from 1.72mm to 3.54mm O.D, so you pick a suitable size for a particular frequency. They are annealed and very nearly keep their shape when wound on a wooden mandrel. Even buying them retail, I get about two 15T to 20T helixes from an $18 pack (about £6 pounds sterling) and many more 4T to 6T ones. Considering they work so well and are easily available, that's not bad. The ex factory price I got is less then half the retail price and they seem as though they would supply me with small quantities, which is a bit unusual.

The retailer also keeps dozens of copper cap and tapered sections like your dual mode feeds. Unfortunately, the factory does not have an agent in UK. He only seems to supply VK, ZL & S. E. Asian countries. I would imagine there are similar sources of tube made in the UK or Europe, so it may be worthwhile looking into. For your information the packs are 150mm diameter and the ODs and lengths of the type they make are as follows:-



PART No.	O.D. (mm)	USEABLE LENGTHS (mm)
SP1	1.72	3660
SP2	2.06	3660
SP3	2.18	3660
SP4	2.16	4270
SP4.5	2.26	4270
SP5	2.58	4270
SP6	2.70	3660
SP6.5	2.82	3660
SP7	2.94	3050
SP8	3.10	3660
SP9	3.06	2750
SP10	3.44	3050
SP11	3.54	3660

YAMSHA

Yet another Mode S Helical aerial

Martyn Kinder G0CZD

Whilst on a trip to the plumbing department of our local B&Q hardware store before Christmas to prepare for a new kitchen sink, I spotted some semi-rigid 42mm plastic waste pipe. A quick calculation.

42mm dia. x pi = 13cms = mode S

and I can fasten my newly converted Drake to this as well. End of design process!

Well, very nearly. Clutching a 2m length of the waste pipe (along with the rest of the copper bends and valves for the kitchen sink) I rushed home and attacked a calculator, the ARRL "Satellite Experimenters Handbook" (1st Edition –1984) and the contents of an extensive junk box. I then put it all away neatly and read pages 6-16 to 6-18 several times over. YAMSHA Mark 1 was conceived. Following many conceptions, nothing else happened for a couple of weeks (except the sink was installed).

One of the claimed benefits of Helical antennas is that they remain quite functional over a large bandwidth, typically 0.8*f*c to 1.2*f*c, with *f*c the "tuned centre frequency". This was just as well. The optimum frequency for an air spaced helical with an internal diameter of 42mm is 2273MHz. A little bit lower than the 2400MHz I really wanted. The remainder of the design parameters were calculated as follows:

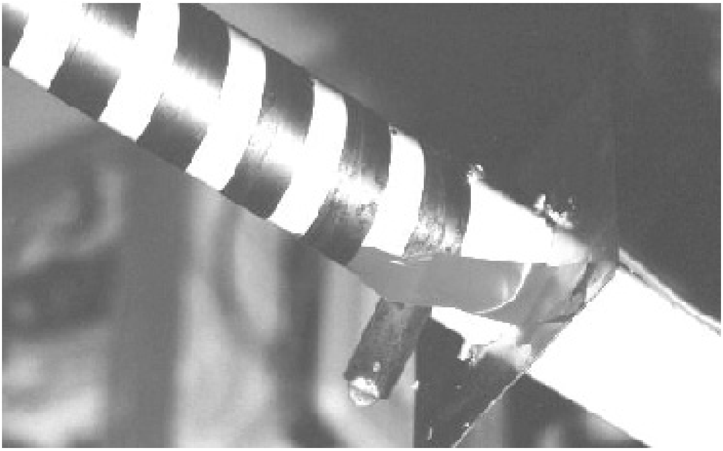
The wavelength of an aerial with 42mm diameter helix is

? = 42 x 3.14 = 131.9mm

To achieve efficient circularity, a pitch angle of 12.5 degrees is required. The spacing between turns can be calculated as S = ? (tan 12.5) = 29.24mm.

Finally the reflector had to be a reasonable size, a minimum side length of 0.6?? is recommended, I settled on a nice round 100mm. These numbers were starting to look silly. There was no way that I (or any other amateur) could measure 20 turns to an accuracy of 1/100mm. A little bit of creative thought was required. But firstly, the reflector had to be manufactured and secured. I decided to use a two hole gold plated SMA socket as the termination/connection point. A 43mm hole was cut in the centre of the 24g brass sheet (available from your hobby shop or much cheaper from a scrap metal merchant), and the SMA connector mounted as near to the edge of the hole as possible, whilst still allowing an SMA plug to be screwed in.

The Plastic pipe was marked up with the reflector 250mm from the mount end. This allows adequate space to mount the Drake and also an aerial mounting clamp. The reflector was located by two short lengths of 3mm brass rod passed through the tube and soldered, one on each side of the reflector. This secures the reflector very solidly. Next, mark on the tube, on the same side as the SMA connector three small pen marks, one 29mm from the SMA, one 88mm from the SMA, the other 585 mm up the tube. Drill a small hole into the tube at this point and cut the tube neatly about 5mm beyond this. You will need a 3metre steel tape measure. Place the tape measure number side down at the first mark and wind two turns around the tube. Adjust the pitch of the measure so that both of the first two marks are on the same edge of the tape. Place some sticky tape over the start of the tape and wind out another 17 turns, keeping the tape absolutely flat against the tube. If you have got it right, you will use all 3 metres of steel tape and will terminate right on the final hole – and with no compound errors. Make sure that you wind the helix in the right direction. Phase 3D will have right hand polarised antennas, this means that in the direction of radiation, the helix should be wound in a clockwise direction. Have a look at the photographs if you are unsure. Now secure the other end of the tape in place with some sticky tape. I used just over 3m of 16g hard drawn copper wire for the "element". Clean the enamel off the end and solder to the protruding pin of the SMA socket. Now keeping the wire tight, twist the aerial one turn until the wire lines up with the correct edge of the tape. Keep winding for the remaining 19 turns using the tape as a guide and then finally push the wire through the hole. Pull tight with a pair of pliers, bend up and cut off any excess leaving about 6mmon the inside of the pipe. Now carefully cut the sticky tape securing the steel tape measure and unwind. Stand back and admire your perfect Helix. The original design had a slightly different feed point from this final aerial design.

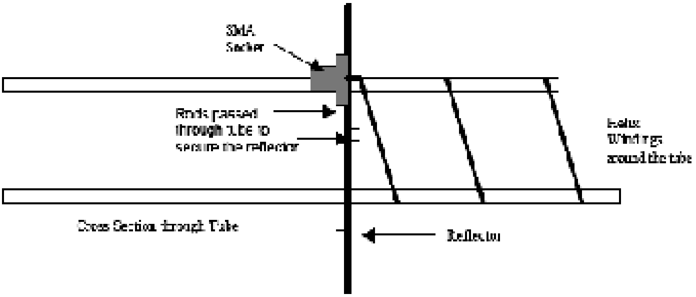


At the time of writing (January 2000), there is only one active satellite with a Mode S beacon. (UO-11 transmits a weak, unmodulated beacon but uses left hand circular polarisation). A second prototype was built using Left Hand circular polarisation.

On the original design (following advice from the "Satellite Experimenters Handbook", the start of the first helix was delayed by 29mm, i.e. the wire came straight from the SMA for 29mm before the windings started. This feed arrangement was designed to improve the match to 50 ohms. I spent around a week trying to hear UO-11 with absolutely no success. The Drake had been tested at the last Adastral park Microwave Round Table in November 1999, and exhibited < 2.0dB noise figure and just less than

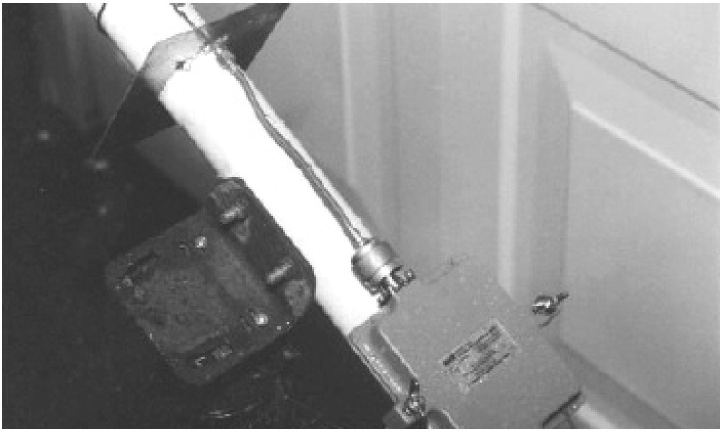
40dB gain. It was working OK!.

Eventually, I decided that the problem was likely to be related to the feed point (no rational reasons – it just didn't look right) of the Helix and so I extended the spiral back all the way to the SMA and tried again. Jackpot! The satellite was picked up on the first pass, with the signal rising to a peak of about 10dB above noise. Very satisfying. Note the photographs show the prototype Right Hand Circular Polarised antenna with the original and inefficient feed.


SMA Socket
Rods passed through tube to secure the reflector
Helix Windings around the tube
Cross Section through Tube
Reflector

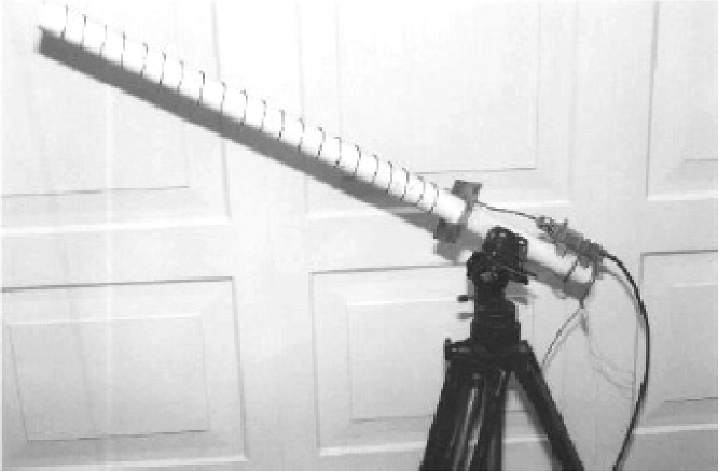
The claimed calculated gain figure for this aerial is about:
10 x log (15 x 20 (turns) x tan(12.5)) = 18.2dBi.
It is generally accepted that these calculated gain figures are in the order of 1 to 2dB too high. (However read Kent Britain's article – no guarantees implied!) 3dB beamwidth is about 30 degrees.

A few other constructional points. The aerial is attached to an 'L' shaped bracket using two 10mm diameter bolts. These were not tightened unduly, but the nuts were secured with Loctite and the bolts "Araldited" into place. A standard ''U' bolt arrangement can be used to secure the aerial either to a vertical or horizontal pole. The size of the aerial is comparatively small and is unlikely to be unduly stressed in high winds. The Drake is attached to the SMA via a 150mm length of UT141 semi rigid coax with the other end terminated in an N plug. When it eventually goes up on



the mast, the connectors will be weatherproofed, the rest of the unit including the Drake will remain out in the Elements.

Postscript: There has been some discussion on the AMSAT-BB mailing list about the feed impedance of helical aerials. The nominal VSWR of this aerial is approximately 3:1. Several suggestions for improving the match have been suggested which will be subject to further tests and experiments. This high SWR does not appear to affect the stability of the Drake 2880, however, some preamplifiers may be inclined to take off with this level of mismatch.



Dishes and feeds

Is Your Dish Doing What You Think It Is?

Richard T. Nadle, K2RIW

INTRODUCTION --
In his 4/15/03 submittal, Kent, WA5VJB quoted the Johnson & Jasik "**Antenna Engineering Handbook", 2nd edition, 1984,** page 36-9 as saying that the reference #16 article authors (Galindo-Israel, Mittra and Cha) are claiming that, "Their analytical techniques are reported to result in efficiencies in the 80 to 90 percent range". Kent is accurately quoting the Handbook, but there is a problem -- I believe the quote is being taken out of context. I'll supply 3 pieces of evidence of this, and I'll provide a further discussion of possible Dish Efficiencies.

(1) Chapter 36 of the Handbook is devoted to "Earth Station Antennas" that are 60 to 300 wavelengths in diameter. As you are about to find out, some great "efficiency magic" is possible when you have a lot of wavelengths to play with.

(2) The type of antennas being discussed on page 36-9 are Double-offset Geometry dish antennas. This means that the feed horn is offset from the hyperbolic-like sub reflector, and the sub reflector is offset from the parabolic-like main reflector. These are rather complicated geometries that our Microwavers may not be able to use, for some time to come (explained below).

(3) The "Analytical Techniques" being discussed are computer-modelled theoretical dish antennas that are using a Cassegrain-like geometry that employs a "shaped" reflector and sub reflector. This means that the main reflector is an intentionally distorted parabola, and the sub reflector is an intentionally distorted hyperbola. The intentional distortions are being used to simultaneously improve the "Illumination Taper" while still providing essentially no unintended "Phase Taper" The computer-modelling techniques had to use "Diffraction Optimization" techniques for which there is no exact Geometric Optics (GO) solution -- whew!

MAXIMUM DISH EFFICIENCY --
I wish we had access to 90% aperture efficient Dish Antennas, and Horn Antennas. We usually do not, and you will soon see why. "Aperture Efficiency" and Side lobe Levels are often misunderstood, and the detriment they cause is often misquoted. I hope the following material will help.

IT'S ALL IN THE FEED SYSTEM --
The requirements for high efficiency are easy to state, but quite hard to achieve. So, what's required for 100% efficiency -- no Phase Error (no phase taper), and no Amplitude Taper. If my great new "wizz-bang" feed horn system could provide exactly the same number of watts per square cm across my parabolic reflector, no spill-over energy at the edges, and no phase errors, than I'd have what's called a 100% Aperture Efficient dish antenna system.

TRANSMISSION & RECEPTION --
During transmit, almost every possible watt emitted by the antenna would be directed toward the target, and during reception, almost every "possible bit" of the signal hitting the antenna (from the correct direction and polarization) would be absorbed and sent down the feed line -- and it would often peg the S meter. If I had such an antenna, there would be no way that I could improve the transmission or reception,

without increasing the size of the antenna.

SIDE LOBES & SCATTERING -- However, newly-instructed antenna engineers are usually quite surprised to find that a theoretical 100% Aperture Efficient (round dish) antenna will still have 1st side lobes that are about 17.5 dB down, and on reception the antenna still has a considerable "Scattering Area" that will show up on a good Instrumentation Radar.

SIDE LOBE ORIGIN -- The side lobes are the result of the Diffraction Effect, caused by the abrupt fall-off of signal at the edge of the antenna. Simplistically, you could say that Mother Nature doesn't like a sharp discontinuity, and she responds by creating those pesky side lobes (they're almost always there).

SCATTERING ORIGIN -- A Free Space vacuum has an "impedance" of 120*Pi (377 ohms). The 100% Aperture Efficient, highly directional, dish antenna, will absorb the maximum signal if it also presents a matched impedance of 377 ohms. However, now there is a 377 ohm antenna across (in parallel with) the free space impedance of 377 ohms. The new local impedance is now 377/2 = 188.5 ohms. A free-space planar wave entering the vicinity of the antenna will "sense" the change in impedance, and a portion of the wave will become scattered at the impedance discontinuity.

MAKE IT DISAPPEAR -- There are ways to make the antenna side lobes as low as you like, and there are ways to lower the Scattering Area as much as you please. But, all of these techniques must be accompanied by an Aperture Efficiency that is considerably lowered -- you can't have it both ways.

IT'S ALL IN THE SIZE OF IT -- I can almost achieve that nearly ideal Illumination Taper (0.0 dB) and Phase Taper (0.0 degrees). However, to have that much control over the radiation characteristics (primary pattern) of the Feed will require the horn to be made from a very large number of radiating elements that have a carefully-controlled amplitude and phase at each element.

Such a "Cluster Feed" would really be an elaborate Phase Array Antenna, and I don't think you can come close to achieving the desired primary pattern with less than 1,000 elements -- a feed with about 38 wavelengths of diameter (43 inches [1.1 meters] at 10 GHz).

SMALL CASSEGRAINS -- I'm always amused when I encounter an enthusiastic, new Microwaver who becomes enthralled with a small-dish Cassegrain antenna system. Quite often they end up with a sub reflector design of 1.5 inch diameter at 10 GHz, and they start to worry about the accuracy requirement for the Hyperbolic shape.

There is a tried-and-true rule of thumb among informed antenna engineers, "don't even consider a Cassegrain system unless the sub reflector is at least 10 wavelengths in diameter". A 1.5 inch sub reflector at 10 GHz (1.3 wavelengths) is essentially a "Point Source" (almost isotropic). If your eyes functioned at 10 GHz, and you looked at an illuminated 1.5 inch sub reflector, it would look like a fuzzy white dot to you, and you couldn't tell if it was concave, convex, or flat -- they would all look the same. It take a large number of wavelengths of object size before your "vision system" begins to resolve the details. The 2 inch diameter of the sub reflector on my 8 inch Celestron Cassegrain telescope is 100,000 wavelengths in diameter -- that's a real sub reflector that can easily control the pattern it reflects!

CONCLUSION -- Concerning those super Aperture Efficiencies we sometimes hear about, they are possible, but difficult to achieve. And, I doubt we will see many of these in our back yards, unless we are talking about elaborate antenna designs that are hundreds of wavelengths in main dish diameter.

Maximum Dish Efficiency and the Best Antenna Side Lobe Levels

Richard T. Nadle, K2RIW

INTRODUCTION --
What follows is long winded but it is intended as a mini-tutorial that I hope will give some microwavers a better understanding about the highly misunderstood area of Maximizing Gain, Aperture Efficiency, Properly Feeding Parabolas, and the Proper Side lobe Levels that must be present in a properly operating, high efficiency, aperture-type antenna.

THE FORMULA --
The most important factor that determines the achievable Gain of a microwave antenna is it's area. The formula that is the bedrock of the antenna measuring/designing industry and science is:

Gain = (4*Pi*Ae) / (Lambda2)

Where:
Ae = Effective Area, often 55% of the Physical Area
Pi = 3.1416
Lambda = Wavelength in the same units as Ae

GAIN EQUALS AREA --
When you study that formula you can come to an interesting Conclusion; at a fixed frequency everything is a constant except the Ae. Therefore Gain equals a Constant x Area. If you want to double the Gain of your antenna (that's a +3.01 dB Gain increase) you have to double it's effective area.

ILLUMINATION --
All of the above assumes that you are properly illuminating that new area you added. In most Parabolic Dish situations (offset and centre fed) that Gain is maximized when you choose a feed horn that has the -10 dB pattern fall at the edge of the illuminated surface (including the extra path length to the edge). That will usually give you a Dish with an Aperture Efficiency of about 55 - 60%.

100% EFFICIENCY? --
You can almost achieve a 100% Aperture Efficiency. All you have to do is design a feed horn that illuminates every square inch of the dish with the same power and have that power abruptly fall off to zero at the edge of the dish (no spill over). To have that much control of the feeds Primary Pattern will require a properly-fed, Cluster Feed, Phased Array of about 1,000 elements, and that feed assembly will be about 30 wavelengths in diameter. If you are working with a Dish that is 120 wavelengths in diameter, this is almost ‘doable’ !

A REAL DISH --
Since many of our antennas are only 20 wavelengths in diameter, that approach is not practical. You would end up with more gain in the feed horn assembly than in the whole Dish antenna system. You would be better off just aiming the feed at the target and eliminating the Dish reflector.

APERTURE EFFICIENCY --
The subject of Dish aperture efficiency is highly misunderstood. amateurs (and engineers) believe that the lack of 100% Aperture Efficiency, or 100% Main Lobe Efficiency, represents a true Power Loss (it does not), and that the "lost power" is in the side lobes (it is not).

THERE IS NO LOSS --
In a reasonably-constructed 55% aperture efficiency Parabolic Dish antenna system, if you apply 100 watts

to that antenna, 99.9 watts will be radiated into space. Aperture Efficiency (surface efficiency) is a measure of the True Gain of your antenna versus the theoretically achievable Gain of an antenna of equal area. The desirable 100% aperture efficiency will only be achieved when:

1. The complete surface is illuminated with the exact same number of watts per square inch.
2. There is no phase error on any of those square inches -- this means no bumps in the reflector and no feed horn phase errors in the Primary Pattern.
3. And there is no spill-over energy being wasted.

WHAT'S PRACTICAL --
We can either loose a lot of sleep fretting over how you are going to make your aperture efficiency go from 55% up to 65%, or you can simply add another foot to the diameter to the Parabolic Reflector (and properly illuminate it) -- both may yield the same gain increase. The second approach is much faster, cheaper, and practical.

MANY ANTENNAS HAVE 100% ? --
The world is filled up with Parabolic Antennas that have an aperture efficiency of about 98% -- they are called "Diffraction Limited" Telescopes. My 8 inch diameter telescope has about that aperture efficiency. It achieves this because the Parabolic Reflector is 370,000 wavelengths in diameter, and the Feed Horn (the Eye Piece) does create the desirable Primary Pattern (it is 9,000 wavelengths in diameter) that allows it to do that.

SIDELOBES vs. EFFICIENCY --
Here is the real kicker concerning side lobes and Side lobe "wasted" energy. A Diffraction Limited telescope could be described as one where the Parabolic Reflector has about 1/20 wavelength accuracy, and the rest of the optical system is working properly. That telescope could easily have an Aperture Efficiency of 98%. That's the highest Gain you are ever going to get out of that available area. But now, lets see what it is really doing.

THE AIRY DISC --
As all astronomers know, every Diffraction Limited telescope creates a "picture" (the antenna pattern) that contains an Airy Disc. That means that around every star in the image you will see some dim rings (the side lobes). The Airy Disk is present in all diffraction limited optics systems (and in all antenna patterns). A proper Airy Disk does not represent a system error. However, if a system error is present, the Airy Disk will change in a characteristic way that's beautifully pictured in Suiter's book, "Star Testing Astronomical Telescopes: A Manual for Optical Evaluation and Adjustment" by Harold Richard Suiter, $29.95 at Amazon.com.

HOW MUCH POWER IN THOSE SIDELOBES? --
From my Melles Griot "Optics Guide 5" catalogue, in the section entitled Fundamental Optics, they say that the Diffraction Limited Airy disc will have a Central Maximum region relative intensity of 1.0 (that's the antenna's main lobe at bore sight), and 83.8% of the energy is located there. The first ring (I call this the 1st side lobe), will have a relative intensity of 0.0175 (I call this -17.57 dB), and will contain 7.2% of the energy.

The second ring relative intensity will be 0.0042 (I call this -23.77 dB), and will contain 2.8% of the energy. The 3rd ring intensity is 0.0016 (I call this -27.96 dB), containing 1.5% energy.

The 4th ring is 0.0008 (I call this -30.97 dB), containing 1.0% energy, and a bunch more dimmer rings with less and less energy (the remaining 3.7%).

100% APERTURE EFFICIENCY CHARACTERISTICS --
Now let's review those last statements. A Diffraction Limited 100% aperture efficient telescope has 83.8 % of the received energy located in the main lobe, 7.2% of the received energy located in the first side lobe, 2.8% of the received energy is located in the second side lobe, and 1.5% of the received energy is located in the 3rd side lobe, etc. These are the best numbers you are ever going to get from a perfect, round aperture, that is not an infinite number of wavelengths in diameter.

REMOVE THE SIDELOBES, NO WAY! --
There is an amazing number of amateurs and engineers out there who are dreaming about getting rid of ALL of those side lobes and their "wasted" energy. This is a VERY FUTILE EFFORT. When a circular aperture HAS 100% aperture efficiency, it WILL HAVE side lobes that are exactly that strong (-17.57 dB [1st side lobe], -23.77 dB [2nd side lobe], -27.96 dB [3rd side lobe], etc.) and the amount of energy in each of those side lobes WILL BE exactly the numbers indicated (7.2%, 2.8%, 1.5%, etc.).

REAL DESIGNS --
You can definitely design an antenna with weaker side lobes; but it WILL HAVE less Gain. You can design an antenna with stronger side lobes; and it also WILL HAVE less Gain. You can then design a low loss (no pads) circular aperture antenna with exactly those magic side lobe levels; and it will have the MAXIMUM GAIN for that size aperture.

IS THIS REASONABLE? --
Of course this doesn't seem to make sense, but that's the way "Mother Nature" and Diffraction Limiter 100% aperture efficiency antennas (and telescopes) behave. Those side lobes are the result of the abrupt change in the illumination taper at the edge of the aperture -- Mother Nature reacts to them by creating side lobes. You could slowly taper the energy as you approach the edge of the aperture; that will decrease the abruptness of the illumination taper and it will lower the side lobes but the available Gain will decrease when you do this. You can't have it both ways (maximum Gain and no side lobes).

SO LET'S STOP THE INSANITY
It's time we microwavers, amateurs, engineers, and interested scientists stopped seeking Maximum Gain antennas that have miniscule side lobes; it isn't going to happen! At least I can say, it's not going to happen in THIS universe, that operates with THIS SET of the Laws of Physics that determine our antenna patterns by using what the mathematicians call Window Functions -- that's the way you feed an aperture.

THE YAGI CONNECTION -- A well-tuned, long Yagi antenna has a nearly circular aperture with a nearly uniform aperture distribution. It is interesting to note that such a Yagi usually has a set of side lobes that are very nearly -17.5, -23.8, -27.9, and -30.9 dB. I think we have been looking at the Yagi antenna's "Airy disc" for a long time, we just didn't give that name.

DISH COMPARISON -- A well tuned Parabolic Dish antenna has weaker side lobes than these, simply because the best available feed horns need to use an Amplitude Taper of -10 dB at the Dish perimeter.

GOOD OPTICS BOOKS -- For those microwavers who wish to dig deeper and try to understand this material I recommend reading some of the better optics books. I soon recognized that the guys who have gotten the subject of High Aperture Efficiency down to a science are the optics people. They can easily do this because their "parabolic antennas" frequently are more than 100,000 wavelengths in diameter. Their "feed horn" is called the eyepiece. Their books can give us a lot of insight into what is really achievable with our microwave antennas.

THE REFERENCE --
Here is what I believe is one of the best books on optics. It's modern, well illustrated with computer-generated graphics and photos, and it's in it's 3rd edition: Eugene Hecht, "Optics", Addison-Wesley, 3rd edition, 1998. It's much nicer than the classic, Born and Wolfe, "Principles of Optics", Cambridge University Press, seventh edition, 1999.

AIRY DISC DEFINITION --
Chapter 5, page 228 of Hecht says: "Because an instrument can only collect a portion of the incident wave front to be reformed into an image, there will always be diffraction: the light will deviate from straight-line propagation and spread out somewhat in the image plane. When an optical system with a circular aperture receives plane waves, rather than there being an image "point", the light actually spreads out into a tiny circular spot (called the Airy disc, containing about 84% of the energy), surrounded by very faint rings. The radius of the Airy disc determines the overlapping of neighbouring images and therefore the resolution. That's why an imaging system that is as perfect as possible is referred to as Diffraction Limited. For a perfect instrument, the ideal theoretical angular resolution is given by the radius of the Airy disc, which is [1.22 x Lambda / D] radians (this is the Raleigh criteria). Another way to present the angular resolution is [2.52 x 105 x Lambda / D] arc-seconds." I added the parenthesis.

Dish Under-illumination, F/D, G/T, Phase Centre and Relate Topics

Richard T. Nadle, K2RIW

Introduction --
I have had a rather close and hands-on relationship with Parabolic Dish antennas since 1968, when I scratch-built my first 12 footer -- "A 12 Foot Stressed Parabolic Dish," (QST, August, 1972, page 16 and on the cover, by K2RIW)

On the WA1MBA Internet Microwave Reflector, Zack, W1VT, asked a simple but provocative question about under illuminating a dish as a method of beam broadening. I call this a "Zoom Control." I have used a Zoom Control for four and a half years, with great pleasure, on my 432MHz array of 16 Yagis (19 elements each = 304 elements total) on a 100 foot tower. The use of electronically selectable beamwidth broadening during a contest operation can greatly increase the fun and eliminate many of the azimuthal ambiguities. Zack's fascinating question kicked off a series of 19 Microwave Reflector responses by K2TXB, G4BRK, K5TR, WA5VJB, W0EOM, KD7TS, KJ4SO, W6CWN, K0CQ, WA2SAY, W7CS, and AL7EB, in that order! Each of the responses contained vital "pieces of the puzzle" but few of the responses would give a newbie a "warm feeling" for what is happening within that modified Parabolic Antenna.

Arguably, the most important and popular type of Microwave antenna is a Parabola. We all want more, new and skilled Microwave operators -- so as to preserve our valuable spectrum, as well as other reasons. For this to happen we must remove some of the mystery about this valuable antenna type. Once learned, many new and crafty operators will "see" new antenna possibilities within every hardware store (such as Home Depot). This memo is my attempt to fill some of that void. Please read also, W1GHZ's On-Line Microwave Antenna Book. The following seven sections try to explain:
(I) The Focusing Action of a Parabola (and its cousins, the Offset Parabola, Ellipse, and Circle),
(II) The two meanings of The Parent Parabola,
(III) The effects of Axial Horn Movement (Prime Fed Dish),
(IV) Axial Horn Movement (Offset Dish),
(V) Non-Axial Horn Motion (two types),
(VI) Gain Maximization
(VII) The W2IMU Horn Modification Problem.

(I) Focusing RF Energy -- During reception, every antenna we use has a certain "capture area" wherein the antenna gathers and focuses the RF energy onto a "driven element," "feed horn," or similar structure that supplies the "gathered RF energy" to the transmission line. In the case of a Parabola, that "gathering," and "Focusing" (if done efficiently) involves at least TWO coherent functions --
(1) having the correct surface orientation (angle of incidence equals angle of reflection), and
(2) obeying the correct phase length (path length). If used correctly, the so-called "conic sections" do this very well.

I'll first give three conic section examples (ellipse, circle, and parabola) to illustrate this, as a kind of thought experiment:

Example (1) Ellipse -- Assume I have a (slightly isotropic) point source RF emitter that's located at the transmitter site, in my back yard, and I wish to capture all it's output at a receiving site, that's 10 feet away. There is one surface that does this perfectly, it's an ellipsoidal surface; it would look like a large, reflective, egg, that's placed around the two sites. That's a three dimensional (3D) ellipse, or an ellipse of revolution about the major axis. I'll boil this down to a 2D discussion.

Assume I have placed a vertical sheet of metal through the two sites and I'll analyse the shape of the ellipsoid that touches the sheet (it's a 2D ellipse). If the ellipse was placed at the coordinate centre, the equation of the 2D elliptical reflector would be (X/a)^2 + (Y/b)^2 = 1, where a and b are constants that define the shape and size of the ellipse.

The ellipse has two foci (focuses); each one is located at one of the sites (transmission and reception). The ellipse has a property called eccentricity, (e = 1 - (b/a)^2), which is less than 1 and somewhat proportional to the distance between the foci.

The magic of the elliptical surface is that it meets the TWO conditions --
(1) every square inch has
the proper surface orientation to reflect the energy in the correct direction, AND
(2) every possible path taken from the transmitter site to the receiver site (with one bounce) has exactly the same path length. That means that the ellipse (ellipsoid) will gather ALL the transmitted energy (from all of 3D space) and focus it (all in phase) at the receiver site. To me, that's kinda neat. By the way, ellipsoidal reflecting surfaces have often been used to focus almost all of the exciter energy of a laser pump onto a laser rod.

Example (2) Circle (Sphere) -- Now let me slowly move the transmitter closer to the receiver, while continuously changing the reflecting surface, so that I can keep gathering all the RF energy. In the limit, the transmitter and receiver will be collocated and the reflecting surface will become a sphere (a circle on my sheet of metal), where a = b in the ellipse equation -- the eccentricity has gone to zero, and the two foci have moved together.

Example (3) Parabola -- Now let me move the transmitter site off to an infinite distance, while continuously changing the reflecting surface, so as to keep gathering all the RF energy. In the limit, that surface will be a Parabola, which is merely an ellipse where the (a) dimension has gone to infinity in the ellipse equation; the eccentricity is 1.0. With a little rearrangement, the equation takes on the familiar form, Y^2 = 4(f)X , where f equals the focal length and Y is the dish radial dimension. Notice that a parabola (for an infinite distance) and an ellipse (for a close distance) are related, but slightly different curves. This may give you some feeling for why there is such a thing as a "Near Field Range" for a parabola (R=2D^2/lambda), where D is the dish diameter and R is the range. If you attempt to make Parabolic Antenna measurements within that range (or closer), the surface shape is sufficiently far from the correct elliptical shape that the RF energy starts focusing slightly out of phase and you start getting a noticeable degradations in pattern and gain. The first sign of this is that the
first null in the antenna pattern (between the main lobe and the first side lobe) disappears.

Moving the feed focal length further away from the dish gives you a slight improvement in the Near-Field Focusing errors, but it does not accomplish a complete correction -- only reshaping the reflector into an ellipsoid would do a perfect job.

(II) The "Parent" Parabola -- This phrase has at least two meanings:

Meaning (1) -- Notice that the basic equation of a parabola always goes to infinity in the multiple directions (X and Y). When we decide to build a Parabolic Dish Antenna, we are deciding to build a reasonable portion of that complete Parabolic curve. By choosing the focal length (f) of our Parabola we are choosing whether the portion we build will be a deep dish or a shallow one -- each has it's advantages. Selecting the focal length (or the F/D) merely selects the radius of curvature (at the apex, for instance). If we cut away some of the reflector, or electrically do a similar function (by under-illuminating), we are not changing the true focal length (radius of curvature) of the Parent Parabola.

In all cases, a well-constructed Parabolic Reflector has ONE FOCUS for all frequencies and dish diameters (when using the same Parent Equation [focal length]). Many amateurs erroneously think that the focal length changes with the frequency or the portion of the surface that is constructed (or illuminated).

Meaning (2), Offset Reflectors -- When we construct a round shaped Parabolic Reflector with the Parabola Apex in the centre, this too can be called the "Parent Parabola." We can then choose to "cut away" an off-centred portion of the reflector, leaving a round (or slightly oval) "Offspring" that includes the apex and one side of the perimeter. That new surface is an Offset Parabolic Reflector. Notice that the Offspring now has a non-symmetric surface that's more curved at the position of the apex of the Parent Parabola (this difference is rather subtle to an unaided eye). Also, there is only one position (spot in 3D space) where the reflector focus is located; you can not rotate the Offset Reflector and leave the feed horn in the same position -- the non-symmetrically curved surface will not focus properly. One of the well kept secrets of almost all the manufactured Offset Fed Parabolic antenna systems, is that the Offspring Reflector includes the Parent Parabola's apex (centre of the original Parent Parabola. The result of this is that you can now easily determine the elevation aiming point. Merely sight from the apex edge of the dish (the part closest to the feed) through the phase centre of the horn; that's the antenna's bore sight. If that turns out to not be true, than most likely you are using the wrong focal position. This became apparent when a number of 10GHz operators started using the same 18" Offset Fed Dish on 24GHz. The dish efficiency was quite low, until they determined the "true" focal position; the shorter wavelengths made this more critical. When they then went back to 10GHz, they discovered a slight increase in efficiency with the "corrected" focal position.

An Offset Fed Parabolic antenna system has the feed horn phase centre placed at the same focal point as the Parent Parabola (it didn't change because of the off-centred cut away). However, the Primary Feed Pattern would now be illuminating areas where the Parent Parabola, was cut away (that would be wasteful). Thus, the feed is re-aimed (but, not translated) at approximately the centre of the remaining Offspring Reflector -- keep the phase centre in the same place. This gives the best (and a higher) efficiency, because now the feed horn is not causing a blockage and the spill over (and feed side lobes) are usually illuminating cold space, and thus not contributing much to the system's noise temperature.

III Shifting The Feed, Forward and Aft (Prime Fed Dish) -- When we move the feed horn along the axis of transmission (in a prime-fed Parabola) we are shifting the focal distance, but we are creating a spherical aberration (a kind of defocusing). Moving the horn outward causes the emitted signal to have a concave wavefront (viewed from a position that's in front of the dish system), and it "Focuses" at a distance closer than infinity (it becomes near sighted). I use the word "Focus" in quotes because, as previously explained (in Section I), this is only a partial phase correction. It makes an improvement for a close emitter, but the technique can only be carried so far. You can't use this technique to get a good focus on a 10GHz 3 foot dish at an emitter distance of 10 feet (mathematics to be supplied later).

Moving the horn inward of the calculated focal distance causes the emitted signal to have a convex wavefront and now the dish is "Focused" beyond infinity (it becomes far sighted). The horn inward moving technique is a coarse method of "Beam Defocusing," or "Beam Broadening" that could be used as a Zoom Control. It causes a more spherical wavefront (lowers the gain), and simultaneously causes the desired under-illuminating function required for a broader beamwidth. There will be some side-lobes developed, but they may be very tolerable; their magnitude is partially dependent on the system's F/D ratio, as well as the Feed Horn characteristics.

IV Shifting The Feed, Forward and Aft (Offset Fed Dish) -- However, the axial motion (along the bore sight) of the feed horn will have a different effect on an Offset Fed Parabolic system. It will simultaneously cause Squint. That's a fancy way of saying that the bore sight will shift. The direction of shift (squint) is fairly easily predicted (with slight inaccuracy) by knowing that (in general) the angle of incidence equals the angle of reflection.

V Shifting The Feed, Non-Axially (Offset & Non-Offset Fed Dish) -- I believe that there is a way of intentionally de-focusing an Offset Fed Parabolic system by moving the horn inward along the axis of the horn. This will slowly (because of the larger F/D ratio) cause a convex wavefront (front view) and, again, it will cause under-illumination -- both are in the desired direction for beamwidth broadening. There probably are past references on this subject but, even if there aren't any (for an Offset Fed system), microwavers are becoming quite proficient with fancy modelling programs; this would be a great place to do some pattern/gain modelling. Or, simply make a "Leap of Faith" -- go do it (try it, you'll like it)!

For a Prime Focus Dish, there is a considerable amount of beam steering that can take place by only moving (translating) the horn in a transverse manner. In the Radiation Laboratory Series of Books ("Microwave Antenna Theory and Design," #11, McGraw-Hill Book Co., NY, 1949, by S. Silver) the author states that the bore sight can be steered by six beamwidths, before the gain falls off by 1 dB, if a 0.6 F/D reflector is in use. The problem becomes worse for lower F/D ratios. **This suggests that a cluster of almost 6 horns, side by side, is possible if a 0.6 or greater F/D is in use.**

I believe that an Offset Fed Parabolic Antenna system will be even more forgiving, when transverse horn motion is used (in azimuth or elevation). This is because of the large F/D ratios that most of them posses. This suggests that a whole cluster of horns is possible, with a rather small gain sacrifice. The only disadvantage may be that a predictable amount of dish re-orientation will be required when a change in horns is initiated.

VI Gain Maximization --

Step One in the process of maximizing the gain of a parabolic Reflector antenna system is understanding the Geometry. This means:

(A) Is the reflector the correct shape and smooth enough for the wavelength in use? If not, can it be re-shaped; can the dents be hammered out, etc.? The Johnson, "Antenna Engineering Handbook", McGraw-Hill, 1992, has curves that predict the rate of gain fall off versus reflector errors (bumps).

(B) Is the reflector's mesh fine enough for the wavelength in use? Should it be covered over with a finer mesh? Johnson (ibid) Handbook has prediction curves.

(C) Has the proper feed horn focal position been found for the reflecting surface in use? By measuring the diameter (D) and the depth (d), and applying the formula,

F = (D^2)/16d ,

This can be found for a Prime Focus Parabola. For an Offset Fed Parabola, the problem is a little more difficult. W1GHZ's web site will help. Even by pure experimentation, it can be found.

(D) Has the system's F/D been correctly determined (this is required to design/ build the proper feed horn)?

Step Two -- The next step is choosing the best Feed Horn to properly illuminate that Parabolic shaped Reflecting surface -- this is a slight compromise. Do you want the maximum Gain, maximum Efficiency, or best Gain to Temperature (G/T) Ratio?. Most of us want the first two (they're very close). An EMEer will want the best G/T. Well, where does the best Gain come from? If this was the best of all worlds, your Feed Horn design would apply an equal amount of RF energy (in phase) to every square inch (or square cm) of the reflecting surface (including the extra path loss to the dish perimeter). That Primary Feed Pattern would abruptly fall to zero at the edge of the dish -- there would be no spill-over energy. Such a Primary Feed Pattern would yield a reflecting surface with 100% aperture efficiency, and the G/T would be ideal.

Believe it or not, those feed horn characteristics are almost achievable. BUT, to have that high a Directivity and Pattern Control, such a horn (or a cluster of horns in a Phased Array) would be larger than most of the Parabolic Reflectors we have ever used. In a Prime Focus Parabolic System, that ideal horn would block out the whole reflector! Even in an Offset Fed Parabolic system, if your horn has more gain than the reflector, skip the reflector and simply aim the horn at the target!

So, for a bunch of reasons, we choose the best realistic horn we know of (Chaparral or Dual Mode [W2IMU] if your dish is near a 0.6 F/D) and adjust them for either a -10 dB, or -20 dB edge illumination taper. The two horn types (Chaparral and W2IMU) are really a multi-element type of feed that have a very desirable pattern, almost no edge currents outside the horn, almost no side lobes, and they have a nearly constant point source phase centre versus azimuth, elevation and diagonal observation angles.

Bear in mind that a subtle shift in the emitted phase from your horn, as a function of a change in observation angle, causes the same system degradation as if your reflector had a big dent in that area. It's hard to believe that some beautiful-looking Parabolic Dish Antenna systems can have a serious error that an untrained eye can not see.

A lot of experience has shown that the best gain occurs at about a -10 dB edge taper (nearly the best illumination taper), and the best G/T occurs at about a -20 dB edge taper (a kind of under illumination).

The gain changes rather slowly (at first) as the illumination taper is changed. The most sensitive characteristic is the first side lobe levels -- they can be as low as -25 dB (or better) with a -20 dB edge taper.

VII W2IMU Horn Modification Problem -- I have read of enthusiastic builders who changed the length or diameter of the large-diameter section, or the diameter of the smaller section, of the W2IMU Dual Mode Horn, as a way of controlling the horn's beamwidth, so as to adapt it to a dish that has an F/D that's quite far from (the designed) 0.6. This can be a disappointing endeavour. Yes, the Dish is not going to crash and burn but it may have a disappointing efficiency (or pattern) result. What Dr. Dick Turrin, W2IMU, did in that horn design was to set up the two circular waveguide modes (TE11 and TM11) in just the right amplitude ratio and phase relationship that the resultant at the horn throat has virtually no edge currents to cause side lobes and back lobes. This also causes the horn to have a constant point source phase centre versus all observation angles.

If you were to change the 30 degree flare angle or the larger diameter, that would change the amplitude ratio of the higher mode generation (TM11). If you change the diameter or length of the larger section, that will change the phase relationship between the modes, because they have different cut-off wavelengths and thus different phase velocities in the larger diameter section. The length and diameter of this section is really a phase corrector between the two waveguide modes.

I may be overly-conservative, but here is my opinion. If you are very skilled at 3D modelling of higher-order waveguide mode generation techniques, and if you are skilled at calculating Bessel-Neumann and Hankel Functions, than have at it and please let me know about your results. For the rest of us mere mortals, I recommend that you follow one of W2IMU's two designs exactly and only scale every dimension, proportional to your particular wavelength.

I hope these thoughts are helpful to those who were brave enough to read all of this.

Please feel free to correct the errors!

Focusing of Deep Focus Dishes

Paul Gaskin, G8AYY

This note is an attempt to show how accurately a deep parabolic dish needs to be focused.

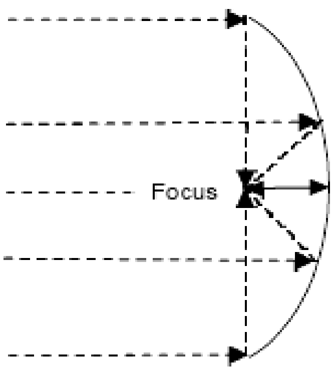
Consider the reception of a plane wave front by a focal plane dish (f/D = 0.25) which is large compared with the wavelength. At the edge of the dish an incoming wave is reflected through 90 degrees which means that the path length to the feed point will not vary significantly with small errors in axial positioning of the feed point.

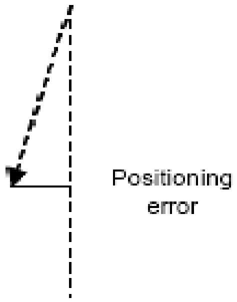
In order to focus, there have to be equal path lengths from the front of the dish to the feed point and an axial positioning error of the feed point will give the maximum phase angle error between the waves from the centre and the edge of the focal plane dish. Loss of gain is related to cosine ((phase angle error)/2) and a phase angle error of over 30 degrees will give a noticeable loss of gain.

At 10 GHz a feed point positioning error of 5mm (1/5 inch) with a deep parabolic dish will have a phase angle error of up to 60 degrees and serious loss of gain which confirms measurements made by other people. If the feed point error was 2.5 mm (1/10 inch) or less then the loss of gain would be negligible.

Measurement of the path lengths from the front of my 45 cm PW dish (f/D = 0.28) to the centre of the disc of the G4ALN 'penny' feed gave a positioning error of only 2 mm (2/25 inch). This dish has a fixed feed point position but works well which confirms the accuracy requirement for the feed point position.

Another 45 cm PW dish was available for direct comparison but did not work so well even after adjustment and requires further investigation. There may be a problem with the dimensions of the G4ALN 'penny' feed.





Large Dish Cassegrain Development For Millimetric Bands & Practical Implementation Using CAD & Spreadsheet

by

Martin Farmer G7MRF

Introduction

The aim of this paper is to make available the design calculations and a spreadsheet program to enable microwavers to design their own hyperbolic sub reflector for a cassegrain feed system as well as a suitable W2IMU feed horn and to explain the implementation of quick band change on my portable set up.

After using flat plate type reflectors on both 24GHz and 47GHz portable systems to good effect, the decision was made to incorporate both of these bands into one box during a major rebuild over the winter months. The decision also included using one dish, with interchangeable bands contained within a quick-change system that could eventually end up with 5.7, 10, 24 & 47GHz as a possible combination on a portable expedition.

The decision to move away from the flat plate type reflectors was influenced by the small improvement in system gain but, more importantly, I wanted to use a W2IMU dual mode feedhorn, using circular waveguide, to feed the new dish (which has a f/D ratio of 0.38).

Looking at the reference books, the optimum f/D for the dual mode feedhorn is 0.8 but it can be used on parabolic dishes with f/D ratios as low as 0.5. The use of a cassegrain system and, in particular, the ability to slightly change values in the calculation makes it possible to modify the f/D ratio of the virtual parabolic dish to suit whatever f/D you require for your feedhorn, as in my particular case from 0.38. If I used a flat plate type reflector, the virtual dish would appear to be 0.38 and not suit the feedhorn type that I wanted to use. For details of the 24GHz and the 47GHz W2IMU dual mode feed horn see **figure 6**.

The original flat plate reflector used on 24GHz was constructed from double-sided PCB material and brazing rods for the supports from the WG20 dish feed. A similar approach was used on my original 47GHz transverter, using a 60mm circular disc reflector. Discussions with Dr Dennis Hawkins of Qpar Angus [1], at the end of one of the UK Microwave Round Tables during 1999, resulted in the exchange of correspondence on the subject of cassegrain feeds.

Armed with this information and calculations from a paper [2] regarding Microwave antennas derived from the cassegrain telescope, we started to lay out, in AutoCAD, the profile from the main dish that was going to be used and, by using geometry, we arrived at a shape for the reflector. This profile from the CAD system was sent to Dr Dennis Hawkins who compared what we had laid out with his commercial antenna software and confirmed it to be ok.

It was decided to go ahead with the manufacture. I found a machine shop that was willing to make the component for me using CNC technology and they had the ability to take the CAD profile of the sub reflector in to their computer system to generate NC code automatically. The end result (shown in **figure 1** below) is the 100mm-diameter sub reflector and support boss used on my new portable system.

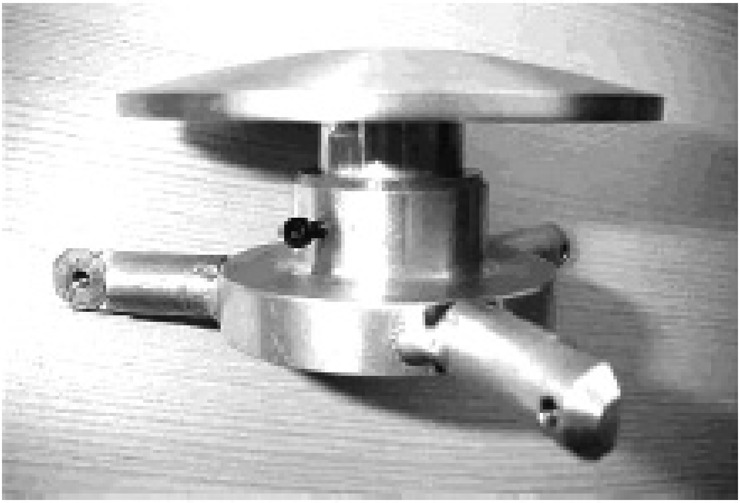


Figure 1: 100mm subreflector and supports

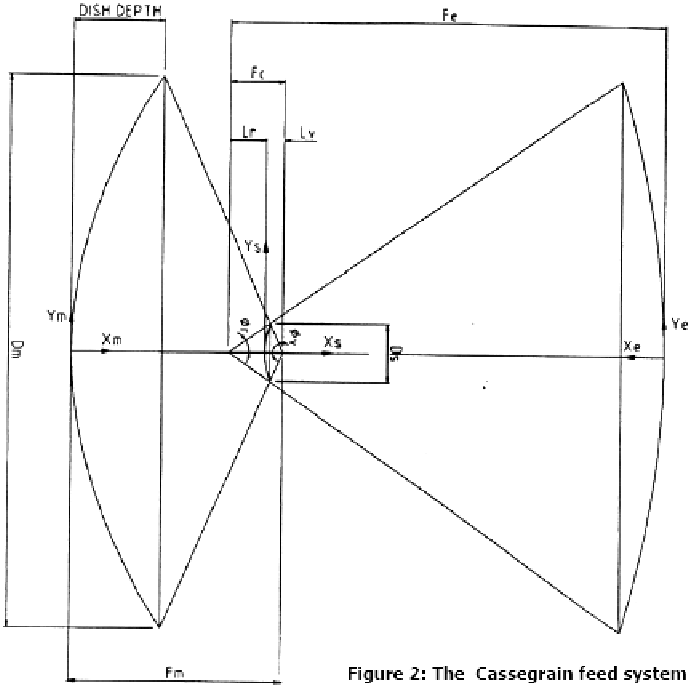


Figure 2: The Cassegrain feed system

Nomenclature

Dm = EFFECTIVE DIAMETER OF CIRCULAR MAIN DISH (TO EDGE RAYS)
Ds = EFFECTIVE DIAMETER OF CIRCULAR SUB DISH (TO EDGE RAYS)
Fm = FOCAL LENGTH OF MAIN DISH
Fc = DISTANCE BETWEEN FOCI OF SUB DISH
Fe = EQUIVALENT FOCAL LENGTH OF CASSEGRAIN SYSTEM
Lv = DISTANCE FROM VIRTUAL FOCUS (OR MAIN DISH FOCUS) TO SUB DISH
Lr = DISTANCE FROM REAL FOCUS (OR FEED) TO SUB DISH.
φv = ANGLE BETWEEN AXIS AND EDGE RAY, AT VIRTUAL FOCUS
φr = ANGLE BETWEEN AXIS AND EDGE RAY, AT REAL FOCUS
E = ECCENTRICITY OF CONIC SECTION.
A = TRANSVERSE HALF-AXIS OF CONIC SECTION.
B = CONJUGATE HALF-AXIS OF CONIC SECTION.
Xm = AXIAL CO-ORDINATES OF MAIN DISH.
Ym = RADIAL CO-ORDINATES OF MAIN DISH.
Xs = AXIAL CO-ORDINATES OF SUB DISH.
Ym = RADIAL CO-ORDINATES OF SUB DISH.
Xe = AXIAL CO-ORDINATES OF VIRTUAL DISH.
Ye = RADIAL CO-ORDINATES OF VIRTUAL DISH

Manual Calculation Method

Here is a worked example based on the following data:

$Dm = 935$ mm

MAIN DISH DEPTH = 152 mm

$Ds = 100$ mm

$Fc = 94$ mm

CALCULATING F/D RATIO

F/D RATIO = Dm/(16 x DISH DEPTH)

$= 935/(16 \times 152)$

$= 935/2432$

$= 0.3845$

CALCULATING MAIN DISH FOCAL LENGTH

$Fm = Dm \times$ F/D RATIO

$= 935 \times 0.3845$

$= 359.5$ mm

CALCULATING ANGLE φv

φv = INV TAN [(Dm ÷ 2) ÷ (Fm - DISH DEPTH)]

= INV TAN [(935 ÷ 2) ÷ (359.5 - 152)]

= INV TAN [(467.5) ÷ (207.5)]

= 66.0659°

CALCULATING ANGLE φr

$(1 \div \tan \varphi v) + (1 \div \tan \varphi r) = 2 (Fc \div Ds)$

$(1 \div \tan 66.0659°) + (1 \div \tan \varphi r) = 2 (94 \div 100)$

$0.4439 + (1 \div \tan \varphi r) = 2 (0.94)$

$(1 \div \tan \varphi r) = 1.88 - 0.4439$

φr = INV TAN (1 ÷ 1.4361)

φr = 34.85°

CALCULATING DISTANCE Lv

$1 - [\sin 1/2 (\varphi v - \varphi r) \div \sin 1/2 (\varphi v + \varphi r)] = 2 (Lv \div Fc)$

$1 - (\sin 1/2 (66.0659 - 34.85) \div \sin 1/2 (66.0659 + 34.85)] = 2 (Lv \div 94)$

$1 - (0.269 \div 0.771) = 2 (Lv \div 94)$

$0.651 = 2 (Lv \div 94)$

$Lv = (0.651 \div 2) \times 94$

$Lv = 30.597$ mm

CALCULATING ECCENTRICITY 'E'

$E = \sin 1/2 (\varphi v + \varphi r) \div \sin 1/2 (\varphi v - (r)$

$E = \sin 1/2(66.0659 + 34.85) \div \sin 1/2 (66.0659 - 34.85)$

$E = \sin 1/2 (100.9159) \div \sin 1/2 (31.2159)$

$E = 0.771 \div 0.269$

$E = 2.866$

CALCULATING VALUE 'A'

$A = Fc \div 2E$

$A = 94 \div 2 \times 2.866$

$A = 16.4$

CALCULATING VALUE 'B'

$B = A \sqrt{E^2 -1}$

$B = 16.4 \sqrt{(2.866)^2 -1}$

$B = 16.4 \sqrt{7.213956}$

$B = 16.4 \times 2.6859$

$B = 44.05$

CALCULATING CONTOUR OF MAIN DISH (PARABOLA)

$Xm = Ym^2 \div 4 (Fm)$

FOR $Ym = 50$

$Xm = (50)^2 \div 4 (359.5)$

$Xm = (50)^2 \div 1438$

$Xm = 2500 \div 1438$

$Xm = 0.017$

REPEAT FOR VARYING VALUES OF Ym TO CALCULATE CORRESPONDING VALUE OF Xm.

EG. FOR Ym = 50 Xm = 1.739

Ym = 100 Xm = 6.954

Ym = 150 Xm = 15.647 Etc

CALCULATING CONTOUR OF SUB DISH (HYPERBOLA)

$Xs = A [\sqrt{1 + (Ys \div B)^2 -1}]$

FOR $Ys = 5$

$Xs = 16.4 [\sqrt{1 + (5 \div 44.05)^2 - 1}]$

$Xs = 16.4 [\sqrt{1 + 0.01288} - 1]$

$Xs = 16.4 (0.006419)$

$Xs = 0.105$

REPEAT FOR VARYING VALUES OF Ys TO CALCULATE CORRESPONDING VALUES Xs.

EG. FOR Ys = 5 Xs = 0.105

Ys = 10 Xs = 0.417

Ys = 15 Xs = 0.925 Etc

CALCULATING EQUIVALENT FOCAL LENGTH Fe

$Fe \div Fm = (E + 1) \div (E -1)$

$Fe \div 359.5 = (2.866 + 1) ((2.866 - 1)$

$Fe \div 359.5 = 3.866 \div 1.866$

$Fe \div 359.5 = 2.072$

$Fe = 2.072 \times 359.5$

$Fe = 744.88$

CALCULATING CONTOUR OF VIRTUAL DISH (PARABOLA)

$Xe = Ye_2 \div 4Fe$
FOR $Ye = 50$
$Xe = (50)_2 \div 4\ (744.88)$
$Xe = 2500 \div 2979.52$
$Xe = 0.839$

REPEAT FOR VARYING VALUES Ye TO CALCULATE CORRESPONDING VALUES OF Xe.

EG. FOR $Ye = 50$ $Xe = 0.839$
$Ye = 100$ $Xe = 3.356$
$Ye = 150$ $Xe = 7.552$ Etc

After going through the above set of calculations a series of X, Y points are obtained. With this the shape of the sub reflector can be laid out onto paper to produce a template. This could then be used to check profile accuracy if manually turning the reflector in a lathe.

After searching the Internet for any programs to calculate the reflector's profile nothing could be found and also not everybody has access to a CAD workstation on their desk so the decision was made to make the calculations available to other amateurs by writing a simple Microsoft Excel spreadsheet to do this task.

Figure 3 shows the layout of the Excel screen. User data input is to the upper left hand side while to the rest are data tables for the sub reflector, actual dish parabola and virtual dish parabola profiles. The dish and virtual dish profiles tables are given to allow the user to draw out the whole scheme if required. I have written an AutoCAD Lisp program to take this data and draw a 1:1 layout within the CAD system. The Excel program is available to download from [3] but we are getting close to the limits of what we as a group are able to do within Excel. Please feel free to play with the routine and comment.

Figures 4 and 5 show the completed portable transverter in use. With using such a large dish, initial alignment when arriving onto the portable site is done by fitting a rifle site into the sub reflector boss on the front of the dish so that it looks back though a hole where the feed fits. By optically aligning the dish onto a visible landmark (180 degrees out) and then working out the beam heading to this point, the compass rose on the neck of the tripod can be altered to suit. After an initial contact, further refinement to the compass can be made.

After working someone with the 24GHz transverter, the head unit can be replaced with the 47GHz transverter, whilst still maintaining the correct beam heading, by the following method ...
I start with standard type die-cast boxes for the equipment and a three-point female conical mounting on the lid of every box that is to be used (see **figure 5**). This is done before any equipment is installed. The position of the waveguide feed is marked using a laser pen innards mounted inside a machined cylinder slid into position where the sub reflector is to be located. The laser light beam projects through a very small hole in the cylinder and the marking of the box entry is then quite straightforward. Mounted on the metal framework that supports the dish are three male conical points that allow the boxes to be repeatedly positioned with good accuracy.

DATA INPUT

Fc (mm)	94.0000
Ds Sub ref dia (mm)	100.0000
Dm Dia main Dish (mm)	935.0000
Fm Focal L Main Dish (mm)	359.4675
Dish Depth	152.0000
Phiv (rad)	1.1531
Phiv (deg)	66.0693
1/tan PhiV	1.4362
Phir (rad)	0.6082
Pgir (deg)	34.8484
Fe/Fm = Magnification	2.0719
Fe	744.7938
Fm/Dm Main Dish	0.3845
Fe/Dm Equivalent Dish	0.7966
Eccentricity of subref	2.8658
Lv (mm)	30.5996
Fc-Lv (mm)	63.4004

Button 10

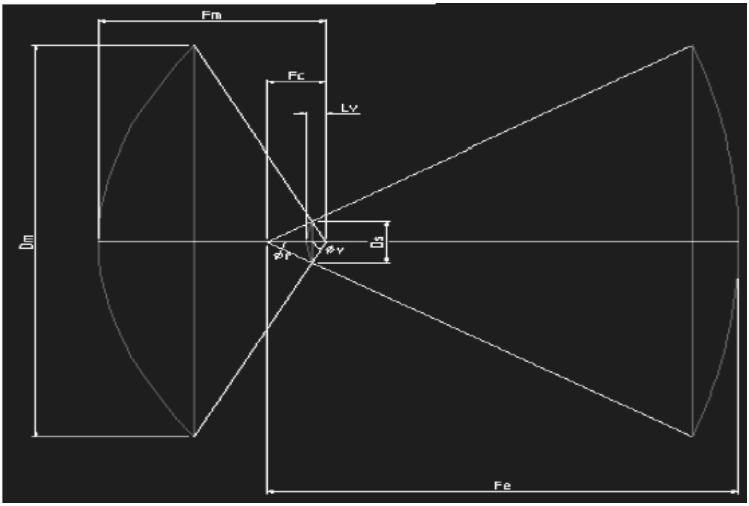
SUB REFLECTOR

A	B	Xs (profile mm)	Ys (radius mm)
16.4004	44.0457	11.3141	-60.0000
16.4004	44.0457	9.8364	-55.0000
16.4004	44.0457	8.4106	-50.0000
16.4004	44.0457	7.0459	-45.0000
16.4004	44.0457	5.7537	-40.0000
16.4004	44.0457	4.5474	-35.0000
16.4004	44.0457	3.4428	-30.0000
16.4004	44.0457	2.4576	-25.0000
16.4004	44.0457	1.6116	-20.0000
16.4004	44.0457	0.9250	-15.0000
16.4004	44.0457	0.4174	-10.0000
16.4004	44.0457	0.1053	-5.0000
16.4004	44.0457	0.0000	0.0000
16.4004	44.0457	0.1053	5.0000
16.4004	44.0457	0.4174	10.0000
16.4004	44.0457	0.9250	15.0000
16.4004	44.0457	1.6116	20.0000
16.4004	44.0457	2.4576	25.0000
16.4004	44.0457	3.4428	30.0000
16.4004	44.0457	4.5474	35.0000
16.4004	44.0457	5.7537	40.0000
16.4004	44.0457	7.0459	45.0000
16.4004	44.0457	8.4106	50.0000
16.4004	44.0457	9.8364	55.0000
16.4004	44.0457	11.3141	60.0000

MAIN DISH PROFILE

Fm Focal L Main Dish (mm)	Xm	Ym
359.4675	156.9161257	-475
359.4675	140.8333095	-450
359.4675	125.6198347	-425
359.4675	111.2757013	-400
359.4675	97.80090938	-375
359.4675	85.19545883	-350
359.4675	73.45934971	-325
359.4675	62.592582	-300
359.4675	52.59515571	-275
359.4675	43.46707083	-250
359.4675	35.20832738	-225
359.4675	27.81892533	-200
359.4675	21.29886471	-175
359.4675	15.6481455	-150
359.4675	10.86676771	-125
359.4675	6.954731333	-100
359.4675	3.912036375	-75
359.4675	1.738682833	-50
359.4675	0.434670708	-25
359.4675	0	0
359.4675	0.434670708	25
359.4675	1.738682833	50
359.4675	3.912036375	75
359.4675	6.954731333	100
359.4675	10.86676771	125
359.4675	15.6481455	150
359.4675	21.29886471	175
359.4675	27.81892533	200
359.4675	35.20832738	225
359.4675	43.46707083	250
359.4675	52.59515571	275
359.4675	62.592582	300
359.4675	73.45934971	325
359.4675	85.19545883	350
359.4675	97.80090938	375
359.4675	111.2757013	400
359.4675	125.6198347	425
359.4675	140.8333095	450
359.4675	156.9161257	475

Figure 3: Excel Spreadsheet



24GHz Scalar Feed and mount for the Direct PC or Prime Start Dishes

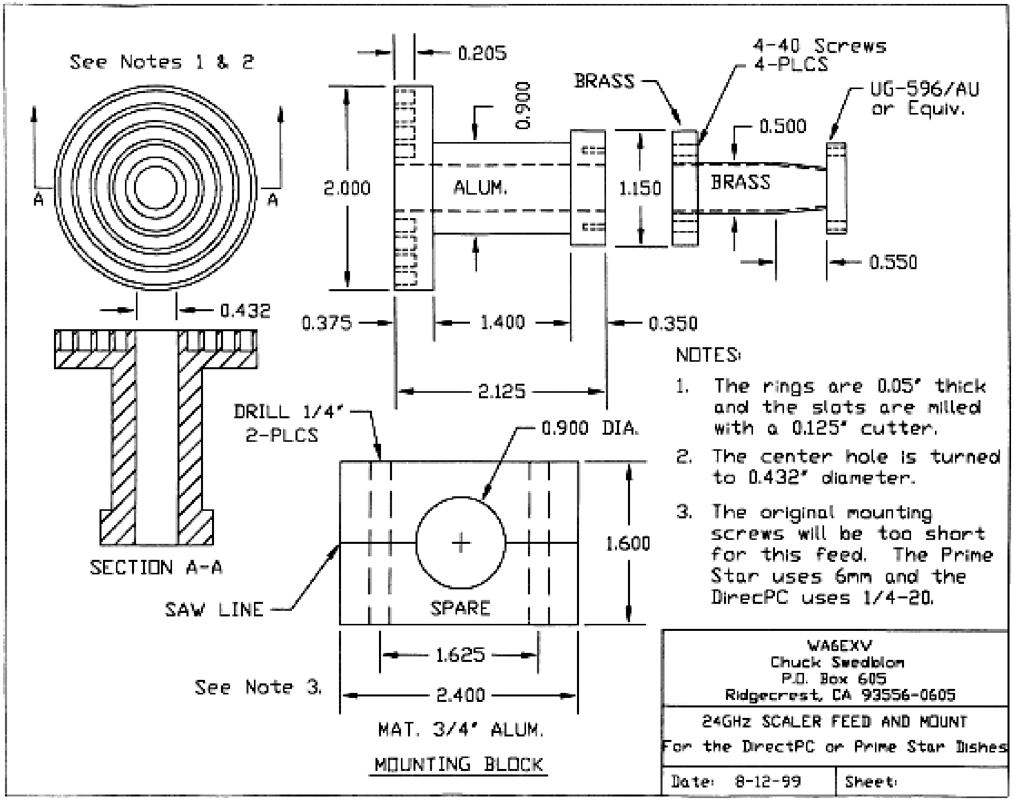
Chuck Swedblom, WA6EXV

Editor's comment: Chuck has very kindly allowed us to reprint this article, originally published in the July 1999 Newsletter of the San Bernardino Microwave Society. While the article refers to two specific brands of dish, the feed is eminently suitable for any prime focus feed with an f/dD of around 0.45 or so.

The existing feed on these two dishes* provided a measured efficiency of 60% at 10.368GHz. I decided to explore the possibility of using the dishes at 24GHz and built this feed to fit in place of the original one. Since my antenna range is too short (92 feet), WA6QYR and I located two sites about a half mile apart on two sides of a small canyon to conduct these test using an antenna of known gain. The results of the tests showed that the DirectPC antenna had an efficency of 59% using this feed at 24.125GHz. The original tests on the Prime Star dish did not perform this good, but the Prime Star dish was used and had been installed here in Ridgecrest for two years or more, whereas the DirectPC dish was brand new. Also there is some doubt in my mind that the Prime Star tests may be flawed. Further tests of the Prime Star will be conducted in the near future.

I built the Scalar Feed, as shown on the following page, since I have the tools and enjoy turning and milling metal. I have also built the scalar section using 1/16" G10 or FR4 PCB material and thin hobby brass strips to form the rings with equal results. In this case the 1/2" brass tubing (circular waveguide) would extend through the centre of the scalar and would have to be long enough to allow for mounting, a total of about 3.5". Also the mounting technique has to be modified such that the centre of the feed will be at the same location as the present one. I used a brass strap soldered to the 1/2" circular waveguide and bent such that the alignment was correct and then drilled two 1/4" holes in the strap to bolt it to the mounting arm of the dish.

The engineering drawing for the feedhorn is on the following page …



NOTES:

1. The rings are 0.05" thick and the slots are milled with a 0.125" cutter.
2. The center hole is turned to 0.432" diameter.
3. The original mounting screws will be too short for this feed. The Prime Star uses 6mm and the DirecPC uses 1/4-20.

WA6EXV
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Ridgecrest, CA 93556-0605

24GHz SCALER FEED AND MOUNT
For the DirectPC or Prime Star Dishes

Date: 8-12-99
Sheet:

Scalar Feedhorn For 24GHz

by Barrie, VE4MA

(reprinted with the permission of the North Texas Microwave Society)

This versatile scalar ring feed first appeared in the June/July 1999 issue of the NTMS newsletter "Feedpoint" and looks just the job for those 24GHz operators who are using dishes with an f/D of around 0.4 or so. For larger f/Ds, such as those found on offset fed dishes, it is suggested you use a dual mode horn of the W2IMU type.

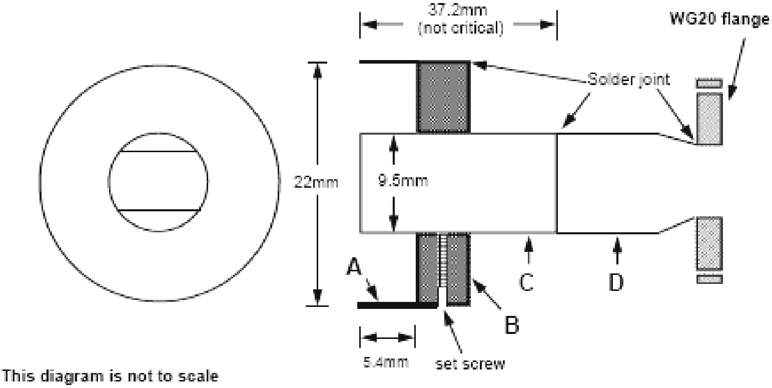
The diagram (fig.1) is largely self explanatory. Part A, the scalar ring, is formed from a pieceof copper waterpipe of an internal diameter that will be a snug fit over Part B, a brass disc(thickness not critical), and to which it is eventually soldered to provide a ring depth of 5.4mm as shown in the diagram. The brass disc is also a snug fit over Part C, a section of brass or copper pipe with an internal diameter of 9.5mm and an external diameter of 11.1mm.

Part D is a soft copper plumbing coupler or pipe of suitable diameter to enable it to be carefully butt jointed and soldered to Part C. Its other end is squeezed and shaped as accurately as possible to fit inside a WG20 flange and then soldered.

For an f/D of 0.4, the open face of the scalar ring should be flush with the open end of Part C. The whole scalar ring assembly (Parts A and B) can be slid back and forth over Part C to adjust for different f/Ds, after which it can be secured with a set screw. Part B is drilled and tapped as shown for the screw.

This design can, of course, be scaled for use on other microwave bands.

figure 1



Using the G3PHO Feedhorn with Coax Feed

by John Quarmby G3XDY

When my satellite system was upgraded to digital I kept the old Amstrad dish with the expectation of putting it to use when I got going on 10GHz. I chose to use the G3PHO "Plumbers Delight" dual mode horn, as this looked like it would illuminate the dish reasonably efficiently, although my dish appears to have a shorter focal length.

The G3PHO design uses waveguide feed with a transition from WG16 to the 22mm copper pipe at the rear of the feed horn. I wanted to build an entirely coax based masthead mounted system, as I already had a Sivers Lab coax relay and all the modules used SMA connectors. It also looked to be easier to engineer the feed to come to the rear of the dish with coax than waveguide, and saved having to find various bits of WG, bends, flanges and transitions. Losses in coax are higher but it looked feasible to keep the cable length short enough to keep within acceptable limits. I ended up using 70cm of Andrews FHJ2 cable for the feeder, with a measured loss of about 0.5dB.

The coax feed design is based closely on the W5LUA dual band feed horn design from Volume 2 of the ARRL UHF/Microwave Projects Manual, optimised for the dimensions of the G3PHO horn.

The diagram, **fig. 1**, on the next page should be self explanatory. Reference to the original article by G3PHO is recommended.

The shorting plate at the end of the 22mm diameter copper pipe is generally to be found in your pocket, and cost the princely sum of £0.01 each! The 22mm pipe is 60mm long and has 44mm exposed beyond the end of the Yorkshire reducer. A piece of brass about 10mm x 10mm x 15mm was filed with a half round file to fit the profile of the 22mm tube. This acts as strain relief for the semi rigid cable emerging from the horn. A 3mm hole was drilled through the middle of this brass block and through one side of the 22mm tube, and they were then temporarily bolted together with a plated steel 3mm screw. The 22mm tube, brass block, and end plate were then silver soldered in place. Silver soldering is more robust than standard tin-lead solder, and because it has a much higher melting point it means that subsequent soft soldering operations on the feed will not cause it to fall apart. When cool the 3mm screw was removed and the assembly cleaned up with wire wool. A 3mm drill passed through the block and tube was gently tapped with a mallet to make an impression on the far wall of the tube, and a 1.6mm hole drilled at this point, then tapped with an M2 thread for the tuning screw. The 3mm hole was opened up to 4mm to accommodate 0.141 inch semi rigid cable. Two examples of the horn were made, one with a cable mounted SMA socket soldered directly to the brass block, the other with a tail of semi rigid cable emerging from the brass block and formed to bring an SMA connector in line with a feeder running along the feed horn mounting arm. The semi rigid cable is soft soldered into the 4mm hole after checking that the feed probe dimensions are correct, with 1.8mm of PTFE dielectric extending into the tube, and

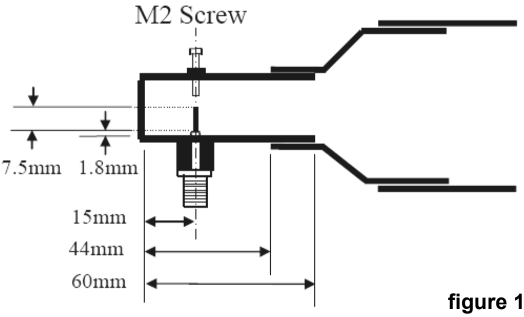


figure 1

7.5mm of inner conductor extending beyond the dielectric. Use plenty of flux to ensure the solder gets right down to the bottom of the hole. The 22mm tube, the Yorkshire reducer, and the 42mm coupler were then soldered in one operation over our domestic gas cooker. Once cool the assembly was cleaned up as far as possible and the outside spray painted to protect it and make it more neighbour friendly.

A 12mm long M2 screw is used to tune the horn, and should have about 5mm penetration into the tube. The two horns I built were tuned up courtesy of Sam G4DDK's excellent test facilities, but can be set up using a microwave power meter and a circulator and tuning for a minimum on the power meter, as per **figure 2** below. No matter how poor the initial match the TX will see 50 ohms thanks to the circulator. Take care not to look into the horn if using more than a few milliwatts for this.

The first feed horn achieved over 25dB return loss over about 200MHz, and could be easily tuned to 10368MHz. The tests were done before soldering the reducer and coupler so that these could also be optimised. The second horn was soldered to the same dimensions as the first before testing so the only adjustment was via the tuning screw, and achieved 20dB return loss at 10368MHz, with best VSWR slightly higher in frequency. I have not made any measurements of the overall dish gain, but the results obtained so far indicate that it seems to work well.

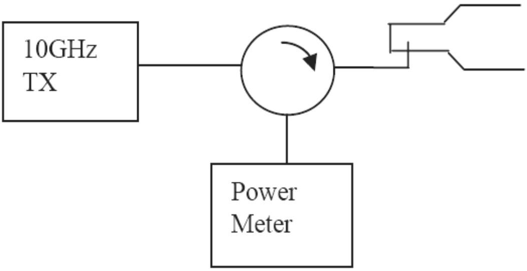


figure 2

3.4GHz Dual Dipole feed

by John Quarmby G3XDY

A description of a dual dipole feed for 2.3GHz suitable for dishes with an F/D ratio of 0.6- 0.7 appeared in Scatterpoint, the UK Microwave Group journal. I have now made a scaled version for 3.4GHz, based on the same principles.

It can be mounted alongside an existing 10GHz feed on an offset fed dish to make a multiband antenna with minimal interaction between the feeds.

On 3.4GHz the beam squint is about 10 degrees off bore sight. This offset has to be taken into account when aligning the dish. The loss due to the offset is about 1dB, so overall gain should be of the order of 24-25dBi for a 60cm dish.

The EIA dual dipole antenna performance as a dish feed is analysed in detail in Paul Wade W1GHZ's excellent online microwave antenna manual at:

http://www.qsl.net/n1bwt/preface.htm

The original 432Mhz feed design is documented in the ARRL UHF/Microwave Experimenters Manual.

The balun design has been changed to make it more readily realised using UT141/RG402 semi-rigid cable, otherwise the design is scaled directly from the original.

The reflector is made from un-etched double sided glass fibre board, with the dipole elements made from 1.4mm diameter copper wire recovered from old cooker mains cable, and the phasing lines made from 1mm diameter copper wire, all soft soldered in place.

I found that the lengths of the dipole elements had to be extended slightly beyond a half wave to get a good match. With 22.5 mm lengths of wire, the return loss was only about 8dB. Extending them to 24mm improved the return loss to 17dB, equivalent to a VSWR of 1.3:1. No doubt this could be improved by further adjustment but the difference in performance would be hard to see.

I measured the return loss using a reflectometer made from 2 pieces of semi-rigid cable, as described on page 10.14 of the Microwave Handbook, but with an interaction length of about 8mm. In addition to adjusting the length of the dipole elements, the match can altered by squeezing the phasing line sections closer or further apart.

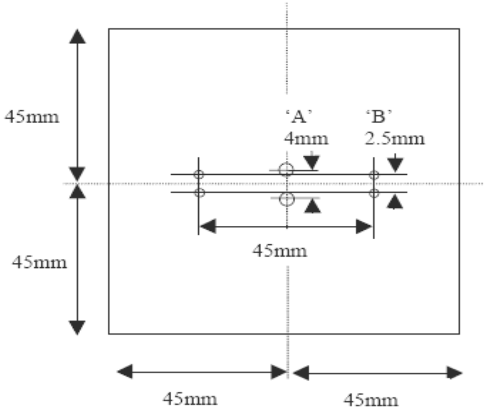
Weatherproofing the feed depends on which way up the dish is mounted. An "upside down" mounting of the dish with the feed support arm at the top means that most of the weather hits the back of the reflector. I just added a piece of Perspex about 60mm x 90mm to the upper edge of the reflector to prevent rain falling onto the upper dipole element. With the dish mounted the other way up then something needs to be mounted over the feed elements to stop water bridging the phasing lines and balun. I found a plastic jar that had contained 500g of mixed nuts that should do the job well.

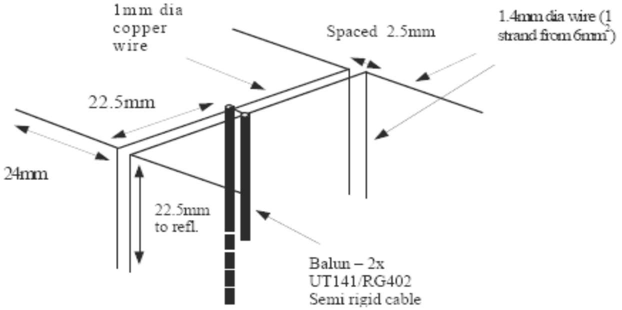
Results have exceeded expectations, with 11 QSOs in the recent UHF contest on one of the least used microwave bands. With the availability of surplus 15W solid state PA s and a new updated transverter design from DB6NT, perhaps we will see a welcome increase in interest in this band.

EIA Dual Dipole dish feed for 3.4GHz

Reflector – 1.6mm double sided copper clad FR4 board 90mm square

Holes A = 9/64" (3.6mm) dia
Holes B = 1.5mm dia




1mm dia copper wire
Spaced 2.5mm
1.4mm dia wire (1 strand from 6mm²)
22.5mm
24mm
22.5mm to refl.
Balun – 2x UT141/RG402 Semi rigid cable

Analysis of the OK1DFC Septum Feed Horn

Paul Wade, W1GHZ

The septum feed1 was described by Zdenek, OK1DFC, at the 10th International EME Conference 2002 in Prague. On-the-air results were promising, but, like any new antenna, there were questions as to how well it really works. Computer simulations suggest that this feed should work well, and also suggest some variations to allow use over a range of dish f/D. The septum polarizer may also be used to generate circular polarization in other feedhorns.

Description

The septum feed as described by OK1DFC is an unflared square horn, or simply a square waveguide, with an internal stepped septum to generate circular polarization. Figure 1 is a cartoon of a septum feed with one wall cut away. The horn is excited by inputs on either side of the septum, with the two sides exciting opposite senses of circular polarization. For EME, this provides separate transmit and receive ports of opposite polarization. The excitation may come from two rectangular waveguides, each matching the dimensions of one-half of the square horn, or from a perpendicular probe on each side of the septum acting as an integral transition from coax to the waveguide. The two methods should provide identical results provided that the waveguide section before the septum is long enough to suppress any spurious modes.

The radiating element, at the aperture, is simply a square horn. Rotated 45 degrees, it is identical to a diagonal horn2; if the diagonal horn is excited with circular polarization, then the radiated pattern should be identical. N7ART has shown3 the diagonal horn to be a good feed, so we might expect the septum feed to be also.

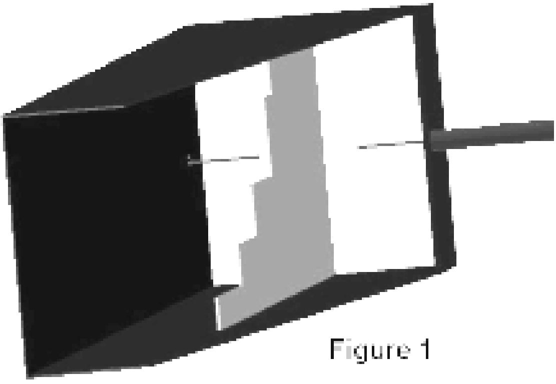


Figure 1

The version described by N7ART used phased crossed dipoles to generate circular polarization. The septum could be a better way to generate circular polarization.

The septum is a bit more complicated. A circularly polarized wave entering the aperture may be considered to have two polarization components with a 90º phase difference, one parallel to the septum and one perpendicular. The parallel component is divided equally by the septum and passes to the two rectangular input waveguides. The cut-off frequency for the perpendicular component is changed by the septum, so that the wavelength for the perpendicular component is shorter. Thus, the electrical length of the septum is longer for the perpendicular component than for the parallel component; if the difference in length is ¼ l, or 90°, then the horizontal and vertical components arrive in phase at the input. The components add together on one side and cancel on the other, depending on the sense of circular polarization, so that the two ports are isolated from each other. In order to achieve the difference in electrical lengths in a reasonable physical distance, the septum polarizer operates near the cut-off wavelength of the waveguides.

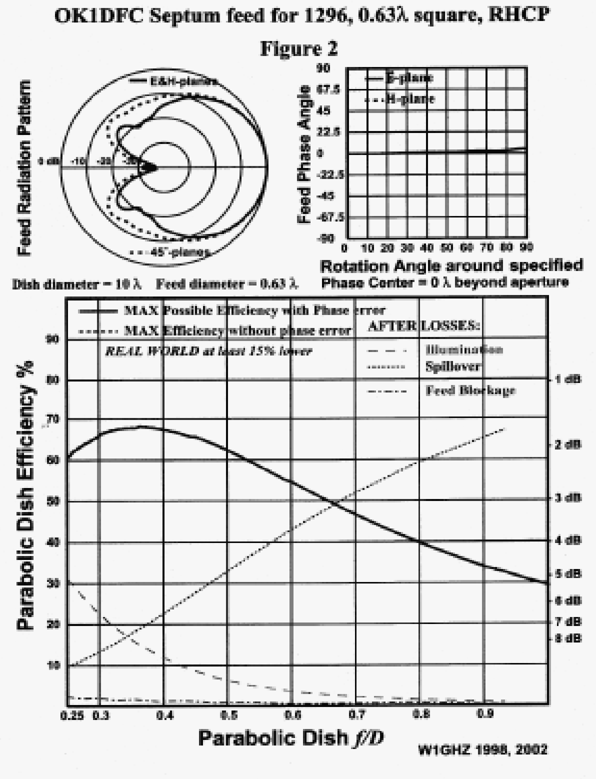

OK1DFC Septum feed for 1296, 0.63λ square, RHCP

Figure 2

Simulations

A septum feed for 1296 MHz with dimensions specified by OK1DFC was simulated using Ansoft HFSS software[4]. The calculated radiation patterns in Figure 2 show the broad illumination expected of a small horn. Like other open waveguide feeds, the rear lobes are relatively large, only about 12 dB down, reducing the calculated efficiency to about 68% with best f/D around 0.35 to 0.4. Patterns for right and left hand circular polarization are pretty much identical. Patterns were calculated for both probe excitation and rectangular waveguide excitation; they were very similar, so the distance from the probe to the septum is adequate.

The circularly polarized pattern of the septum feed, shown in 3D in Figure 3, shows side lobes on the four corners like the diagonal horn, generated as the polarization vector passes through horizontal and vertical polarization in the square horn. The side lobes on the corners reduce the calculated efficiency by perhaps four percentage points compared to a calculation using only the traditional horizontal and vertical pattern cuts.

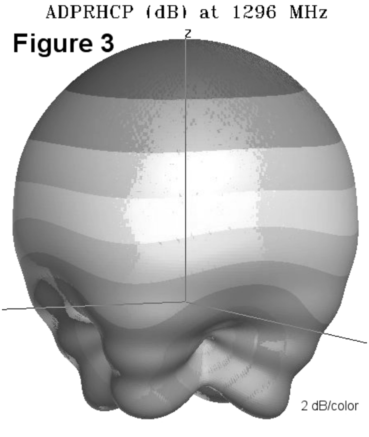


Figure 3

The circular polarization is quite good, with cross polarization about 21 dB down, and the pattern circularity is good. Isolation between the two ports is about 24 dB at 1296 MHz, with reasonable bandwidth, showing good isolation from at least 1.2 to 1.4 GHz. Note that reflection from the parabolic reflector reverses the circular polarization, so that the reflection coming back into the horn will reduce the isolation.

While the calculated efficiency of this feed is not as high as some, the better ones have a larger blockage shadow, so the septum feed may be the best performer on a small dish where circular polarization is required

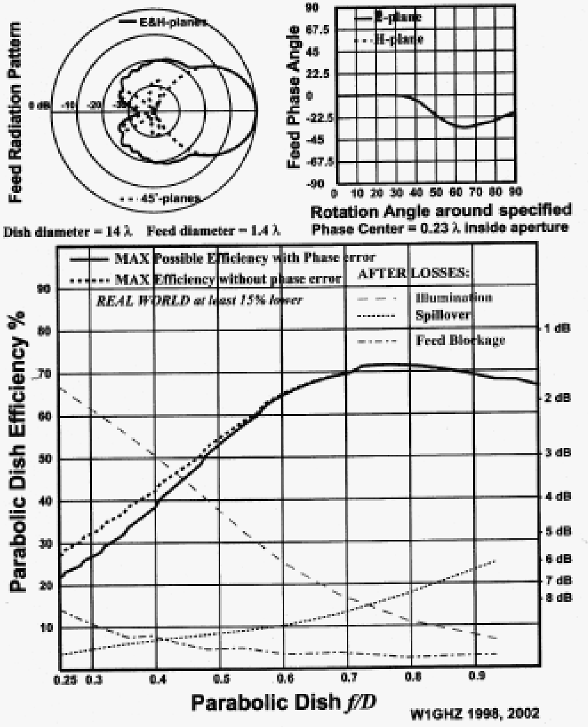
Other f/D dishes

The diagonal horn may be tailored to illuminate a various **f/D** by varying the dimensions of the diagonal section, or by adding a flared section for larger **f/D**.

Since the operation of the septum in generating circular polarization depends on the guide dimensions being close to the cut-off wavelength, the square cross-section is fixed at 0.63l for a given operating frequency.

However, a flare section may be added to increase the aperture size to optimize the horn for any larger **f/D**, so that the septum feed may be used for any dish with **f/D >** 0.3. The flare section is similar to a rectangular waveguide horn, except that it should maintain a square cross-section with a gentle taper to prevent excitation of unwanted modes.

I first tried a adding a flare section with an aperture 1.4l square and a flare angle of 30° (15° half angle on each side of the septum), since this size diagonal horn with linear polarization is a good feed for an offset dish with an equivalent **f/D** around 0.7. With the septum feed generating circular polarization, the calculated efficiency in Figure 4 is high with best **f/D** is around 0.7 to 0.85, suitable for many offset dishes.



OK1DFC Septum feed with flare to 1.4λ square at 1296, RHCP

Figure 4

This horn also had high rear side lobes on the corners, so that the 3D pattern in Figure 5 looks like a rocket with fins. An intermediate size flare, with an aperture 1.1l square, produces the radiation patterns with high calculated efficiency at intermediate f**/D**, best around 0.5 to 0.6, and less pronounced corner lobes.

Both flared septum horns show good isolation and cross-polarization. Since horn beamwidth is inversely related to aperture size, we can choose an appropriate aperture for the flare for any f**/D** by interpolating between the results for the three sizes above, 0.63l square, 1.1l square, and 1.4l square.

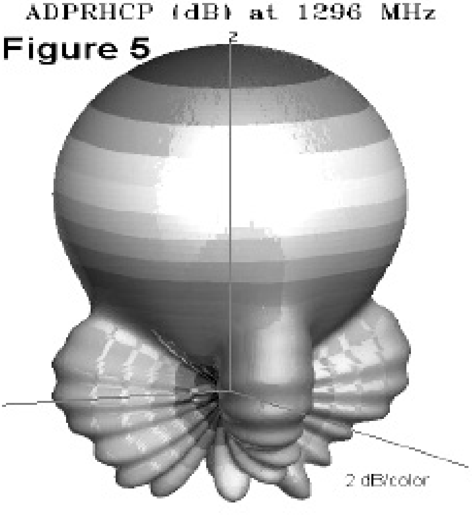
For smaller apertures, the flare angle should be small so that the flare length is reasonably long.

Summary

The septum feeds are impressive — a feedhorn with good circular polarization performance with no adjustments and no phasing losses. The simple square cross-section described by OK1DFC is ideal for low blockage on small deep dishes, while a choke may be added for better performance on larger dishes. A flare section to increase the aperture will better illuminate shallow and offset dishes. The septum polarizer can also be used in cylindrical horns like the VE4MA feed. More information is available on the OK1DFC (www.qsl.net/ok1dfc) and W1GHZ (www.w1ghz.org) web pages.

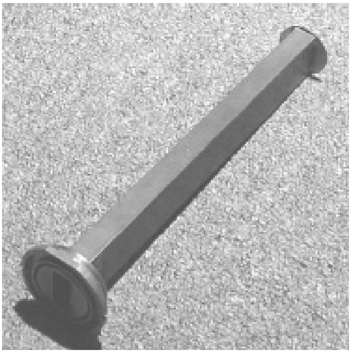
References

1. Zdenek Samek, OK1DFC, "Feed for Parabolic Dish with Circular Polarization," 10th International EME Conference 2002, Prague, 2002. www.qsl.net/ok1dfc
2. A.W. Love, "The Diagonal Horn Antenna," Microwave Journal, March 1962, pp. 117-122. (reprinted in A.W. Love, Electromagnetic Horn Antennas, IEEE, 1976, pp. 189-194.)
3. R. Miller, N7ART, "A 23cm Diagonal Waveguide Feed," DUBUS, 2/1997, pp. 514.
4. www.ansoft.com



G4ALN Waveguide Feed ReVisited

Paul Gaskin, G8AYY



This popular feed was originally described in Radio Communication, October 1976, 'Microwaves' p757 and was intended for used with short focal length dishes with an f/D in the region of 0.25-0.3. The use of an old penny coin for the disc on 10GHz came later. The width of the ë/2 slots was specified as ~ ë/20. It was claimed that the length of the slots, and also the diameter of the disc, were probably not critical within a few per cent and the width of the slots even less so. This seems to overlook the fact the slots are located next to the reflector disc (splash plate) which theoretically has zero electric field strength at its surface in the waveguide. This could make the width of the slot quite critical. Matching screws or a sliding sleeve was suggested to improve the match but no thickness was actually specified for the disc which could be important.

The G4ALN 'penny' feed became popular with the 45 cm PW (Practical Wireless) dish, f/D 0.28, introduced in 1981 for a 10GHz project, and also other prime focus dishes, but did not always work as well as it should have done. I had two different versions of the 'penny' feed in 45cm PW dishes available for tests and found that one version did not work as well as the other with a Solfan head although it had the specified ë/20 slot width and an adjustable feed point position.

This 'penny' feed was modified as part of an investigation into its performance. The 1mm thick brass disc was replaced with a thicker brass disc which was silver plated and the width of the slots was increased. The new disc was soldered to the wave guide, with heavier soldering at top and bottom for strength but fillets of solder were avoided near the slots in the broad faces.

There was an improvement in performance which suggests that the problem could have been caused by the lack of matching screws and or the illumination of the dish.

My version of the 'penny' feed seems to work well, is capable of being sharply focussed and cannot be improved with matching screws. The 2.5 mm thick discs were cut from an old silver plated brass chassis and filed to match an old penny in diameter. The waveguide slots were experimentally increased to 5mm (~ëg/8) width which seems to give good illumination and matching. The clean silver plating on the 31 mm diameter discs probably improves their efficiency as reflectors.

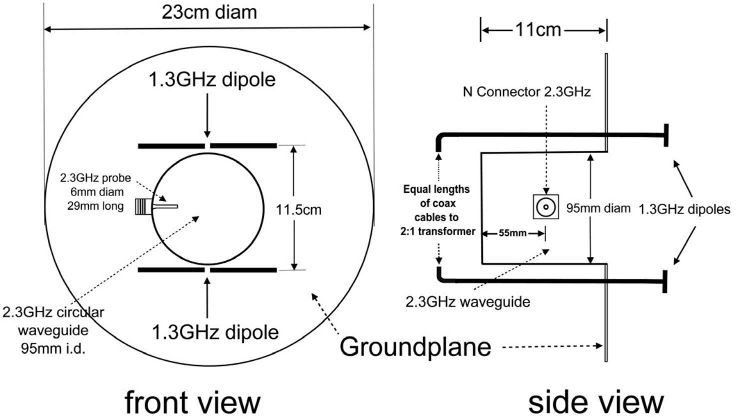
Versions of the G4ALN feed have produced for the 5.7, 24 and 47GHz bands also (reference RSGB Microwave Handbook Volume 3). They all use the ë/20 slot width and it is suggested that it may be worthwhile increasing these slot widths also. Incidentally if anyone is thinking of using an 'English' silver coin as a disc then he should be aware that British 'silver' coins have been made of cupro-nickel alloy since 1947.

A Dual Dish Feed for 23cm and 13cm

by Peter Blair, G3LTF

The diagrams that follow are self-explanatory. A copper or brass disc, 23cm diameter, is soldered to the outer lip of a 95mm i.d. copper or brass tube (or suitable soup can! ...editor). The tube forms a waveguide feed for the 2.3GHz amateur band while the disc becomes the reflector for a pair of 1296MHz dipoles. The rear of the copper tube is, of course, blanked off. A 29mm long probe made from 6mm brass rod or tube is soldered over the end of an N socket centre conductor and the resultant assembly is fixed to the 2.3GHz feed horn so that the probe is 55mm from the horn's back plate and in the horizontal polarised position as shown in the first diagram. That is the 2.3GHz (13cm) section of the dual feed completed.

The 1296MHz feed consists of a pair of stacked dipoles, each made as shown in the second diagram on the next page. They are mounted so that they are in front of the 2.3GHz feed-horn and critically spaced from the circular reflector or ground plane.

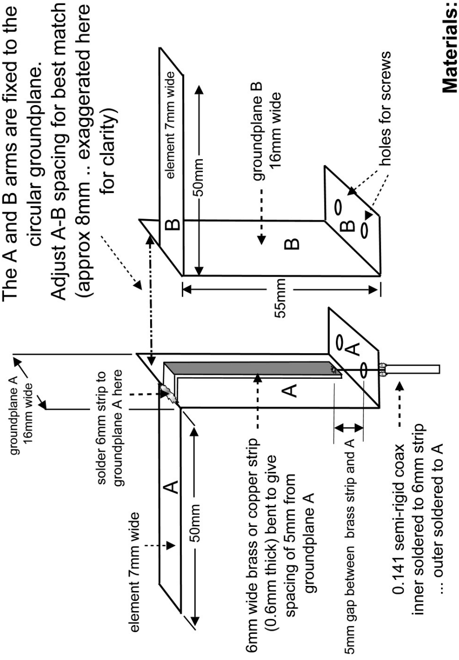


G3LTF Dual Band Feed for 23cm and 13cm

Identical lengths of UT141 semi rigid coax, terminated in sma connectors and fed from the main 1296MHz feedline (e.g. heliax or similar) via a 2:1 transformer (see final diagram), are fixed through the ground plane and rigidly mounted to it. By varying the centre spacing between each arm of each dipole, as shown in the third diagram, a good match can be obtained.

This feed has been very successfully used for several years on a 2m diameter dish whose f/D is 0.5.

The feed can be suitably protected from the weather by the judicious use of plastic boxes and sealing compound.


The A and B arms are fixed to the circular groundplane.
Adjust A-B spacing for best match (approx 8mm .. exaggerated here for clarity)
element 7mm wide
50mm
groundplane B 16mm wide
B
B
B
holes for screws
55mm
groundplane A 16mm wide
solder 6mm strip to groundplane A here
A
A
A
element 7mm wide
50mm
6mm wide brass or copper strip (0.6mm thick) bent to give spacing of 5mm from groundplane A
5mm gap between brass strip and A
0.141 semi-rigid coax inner soldered to 6mm strip ... outer soldered to A

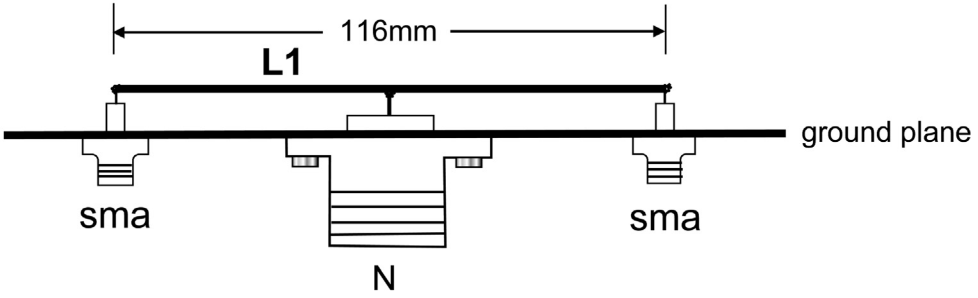
Materials:
Element and groundplane is 0.6mm brass or copper sheet cut and bent as shown in "one piece construction".
Note the vertical sections are 16mm wide and the horizontal dipole sections are 7mm wide

23cm dipole assembly

2:1 transformer

(enclose in suitable waterproof box)

L1 : 13mm wide copper or brass strip spaced 5mm over ground plane



Analysis of the G3LTF Dual Band Feed for 23cm and 13cm

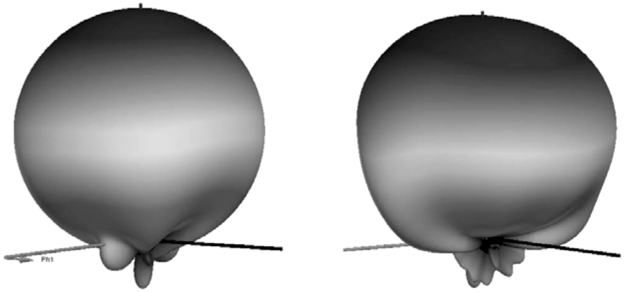
Paul Wade W1GHZ

G3LTF described has described **(1)** a dual band feed for the 23 cm and 13 cm bands. The feed is a combination of two proven designs: an EIA dual-dipole feed for 23 cm, with a cylindrical waveguide horn, or "coffee-can," feed for 13 cm. The dipoles are backed with a 1λ diameter GP (ground plane); the cylinder fits in a hole in the centre of the GP, so that the GP acts as a flange at the aperture of the cylinder. Each dipole is fed with a stripline balun arrangement from a remote power divider, and the cylindrical waveguide is excited by a probe.

Taken separately, each feed is known to work well; radiation patterns and calculated dish efficiency may be found in the W1GHZ Microwave Antenna Book — Online **(2)**. What remains to be seen is how well the combination works. Is the dual band feed as good as the individual feeds, or are there unexpected interactions that spoil the performance?

The radiation patterns were calculated using Ansoft **HFSS** software **(3)** and plotted in 3D in Figure 1. Performance as a dish feed was estimated using my **PHASEPAT** software **(2).** The calculated dish efficiency at 2304 MHz is very good, as shown in Figure 2. Best **f/D is around 0.5, just** right for the two-meter diameter dish at G3LTF. The phase centre is 0.07λ beyond the aperture of the cylindrical horn, or about 9 mm above the GP. Since the phase centre is most critical at the higher frequency, the 1296 MHz dish efficiency in Figure 3 is calculated at the same phase centre (best phase centre at 1296 MHz would be 0.14λ above the GP). Efficiency is also very good, with only a tiny phase error due to the phase centre compromise. Best f/D is again around 0.5, so this is a good feed on both bands for the G3LTF dish.

Figure 1

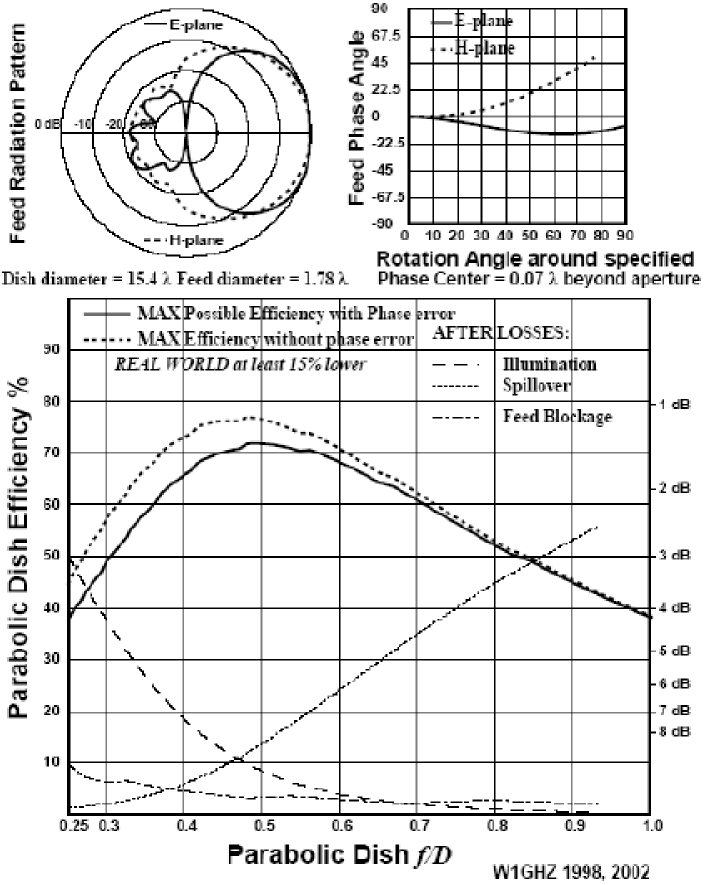


1296 MHz 2304

Since the two dipoles for 23 cm are fed separately, it is important to keep them in phase, by using a good power splitter and identical cables. Figure 4 illustrates the radiation pattern resulting from feeding the dipoles 180º out of phase — the beam splits into two lobes with a null in the centre.

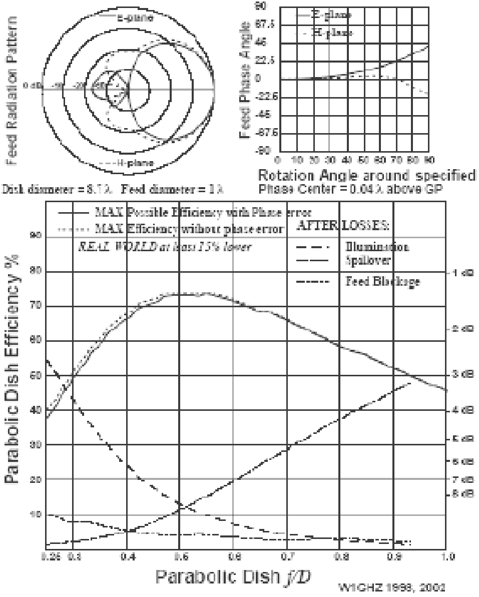
G3LTF L&S Dual Band Feed at 2304 MHz

Figure 2



G3LTF L&S Dual Band Feed at 1296 MHz, 13cm Phase Center

Figure 3



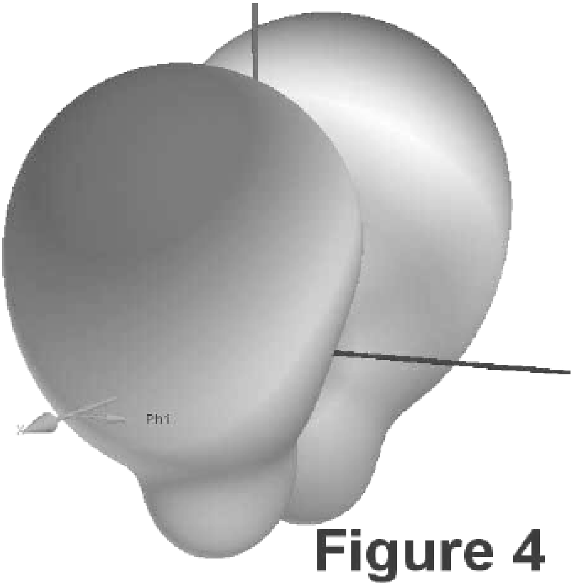


Figure 4

Isolation is always a problem with multi-band feeds. The 13cm cylindrical waveguide is beyond cut-off at 23cm, so isolation at 23cm is better than 30 dB. At the higher frequency, there is nothing to isolate the two feed except spacing, so the isolation at 13cm is on the order of 20 dB. At any reasonable power level, additional protection will be necessary.

References:

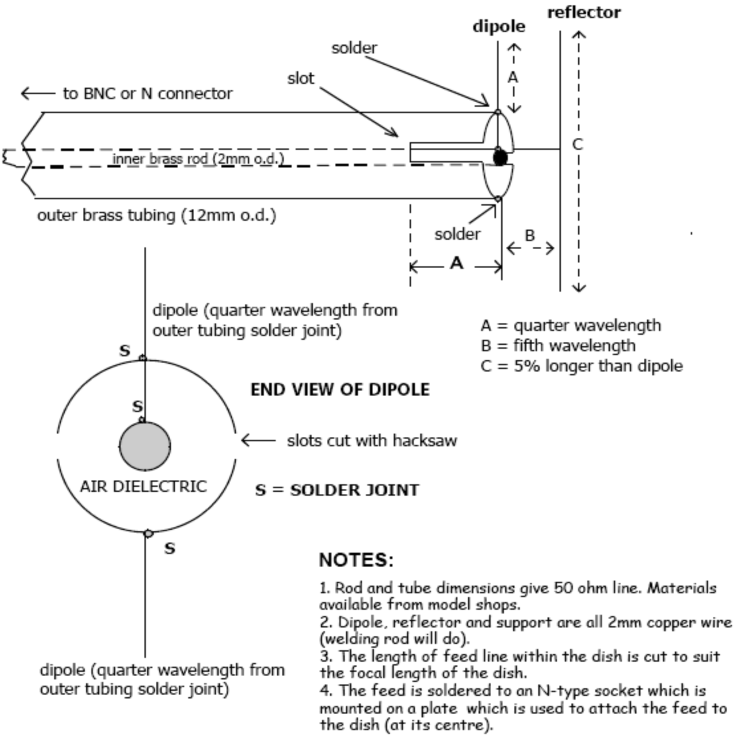
1. Peter Blair, G3LTF, "A Dual Dish Feed for 23cm and 13cm," RSGB Microwave Newsletter, March 2004, pp. 7-8.
2. www.w1ghz.org
3. www.ansoft.com

A Simple Dish Feed For The Middle Microwave Bands

by Mike Parkin , G0JMI

The feed shown in Fig.1 is used at this station on the 6, 9 and 13cm bands. Its most useful feature is its simplicity. The dipole driven element is merely soldered to the supporting brass tube, which in turn forms the feed line. The quarter wave slots form stubs that ensure the dipole is effectively mounted at the end of a quarter wave line.

Figure 1: Plan view showing tubular feedline and wire elements


reflector
dipole
solder
slot
← to BNC or N connector
A
C
inner brass rod (2mm o.d.)
outer brass tubing (12mm o.d.)
solder
B
A
dipole (quarter wavelength from outer tubing solder joint)
S
S
END VIEW OF DIPOLE
← slots cut with hacksaw
AIR DIELECTRIC
S = SOLDER JOINT
S
dipole (quarter wavelength from outer tubing solder joint)

A = quarter wavelength
B = fifth wavelength
C = 5% longer than dipole

NOTES:

1. Rod and tube dimensions give 50 ohm line. Materials available from model shops.
2. Dipole, reflector and support are all 2mm copper wire (welding rod will do).
3. The length of feed line within the dish is cut to suit the focal length of the dish.
4. The feed is soldered to an N-type socket which is mounted on a plate which is used to attach the feed to the dish (at its centre).

Improving the Dual Band 10 & 24GHz Feed Horn for Offset Dishes

Paul Wade, W1GHZ

Operating multiple bands in the 10GHz and Up contest is difficult with separate antennas … after locating a station on 10GHz and peaking the dish, we must start over on a higher band, usually with a narrower beamwidth. Using a dual-band feedhorn for 10 and 24GHz would very attractive; the dish may first be pointed and peaked up on 10GHz, then switched over to 24GHz with no repositioning required.

At Microwave Update 2001, Gary, AD6FP, and Lars, AA6IW, described (**1**) a dual-band 10 and 24GHz feedhorn for shallow and offset dishes. The design was based on previous work of W5LUA (**2**) and W5ZN (**3,4**) to develop a dual-band feedhorn more suitable for conventional deep dishes.

With the offset dish, we have a distinct advantage – the equipment may be located very near the feedhorn without being in the radiation pattern, minimizing the large feedline losses at the higher microwave frequencies without decreasing gain. Other advantages include higher efficiency feedhorns, less critical focusing, and the ready availability of modest-sized DSS dishes with good surface accuracy.

Gary and Lars included computer simulated radiation pattern plots which look like potentially good feeds, but did not do dish efficiency calculations. However, they did include more important results – sun noise measurements and on the air performance! To calculate efficiencies, I took the published dimensions and re-simulated. The results were good, so I wanted to make a feed.

Dual band feedhorn operation

The basis of the dual-band feedhorn design is the W2IMU dual-mode feedhorn (**5**), dimensioned to feed an offset dish at 24GHz and excited from the rear with a circular waveguide section. For 10GHz, an excitation probe fed by an SMA connector is added on the side

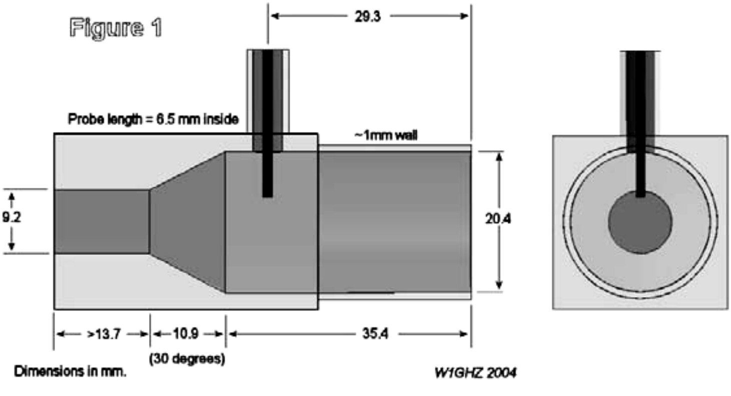


Figure 1

of the output section of the dual-mode horn. The tapered section of the dual-mode horn acts as a closed end at 10GHz, so that the output section behaves like a simple "coffee-can" feed at 10GHz. **Figure 1** is a sketch with the dimensions I used.

While the 24GHz dual-mode horn has a pattern suitable for an offset dish, the simple 10GHz horn has a much broader pattern**,** better suited to a deep dish, so it would have a lot of spill over feeding an offset dish. AD6FP improved the 10GHz performance by adding a conical horn to narrow the beam, and AA6IW enhanced it further by using a corrugated horn.

The dual-mode horn is intended to eliminate edge currents in the rim of the horn, so the addition of the conical horn outside the rim has a much smaller effect at 24GHz. By varying the horn dimensions, it might be possible to make the patterns and efficiencies very close on the two bands. I had four different corrugated horns on hand, so I tried simulating with each of them. Results were promising, so I bored out the circular waveguide end of each horn on my lathe so that it could be slipped over the end of the dual-band horn.

Construction

Gary and Lars built their horns with copper plumbing and hobby brass, soft-soldered together. I tried this construction, but wasn't happy with the dimensional accuracy, and it certainly didn't feel robust enough for rover operation. Then I experimented with turning the tapered section out of solid brass but found it difficult to get the taper right. Finally, I ran out of time before the 2003 contest and simply used a 25dB horn on 24GHz; at least it was easy to point.

Last fall, I was browsing through a tool catalogue from MSC (**6**) and found some 60º countersinks (normal is 82º or more). This would make a 30º flare angle for the tapered section, while Gary and Lars used my HDL_ANT program to calculate a 27.8º taper angle. The 60º countersink would be an easy way to machine a 30º taper, but is 30º close enough to 27.8º ?

I simulated the horn with 30º taper using Ansoft HFSS software (**7**). The rear lobe is a couple of dB worse than with the 27.8º taper, but the calculated efficiency of 76% is close, and the best f/D is about 0.7, just right for a DSS offset dish. At 10GHz, calculated efficiency is still good, about 70%, but best f/D is about 0.38. At the 0.7 f/D needed for an offset dish, efficiency is down to about 47%, or nearly 2dB worse. A few additional trials at 24GHz suggested that a slightly longer output section might be a little better, if the countersink were long enough, but the improvement was not significant.

Another problem is that the nearest countersink size is 3/4 inch, or 19.05mm. I simulated with the inner diameter of the output section reduced from 20.4 mm. This change did not work well at all — the larger diameter is required.

The other mechanical problem is robust feedline attachment: WR-42 waveguide for 24GHz and an SMA connector for 10 GHz. Using brass or copper for the feedhorn would allow soldering, but both are heavy and expensive. Turning the feed from aluminium rod was the best choice, but the size would have to be large enough for the WR-42 hole pattern: 7/8" square, or 1.25 inch diameter. I found that 1" square aluminium was readily available in short lengths, so I ordered some, along with a 3/4" 60º countersink. After a couple of hours with the lathe, my first attempt is shown **in Figure 2.** The machining was possible, but the 3/4" countersink diameter is smaller than the 20.4 mm inner diameter that a small shoulder was left. I fiddled with a

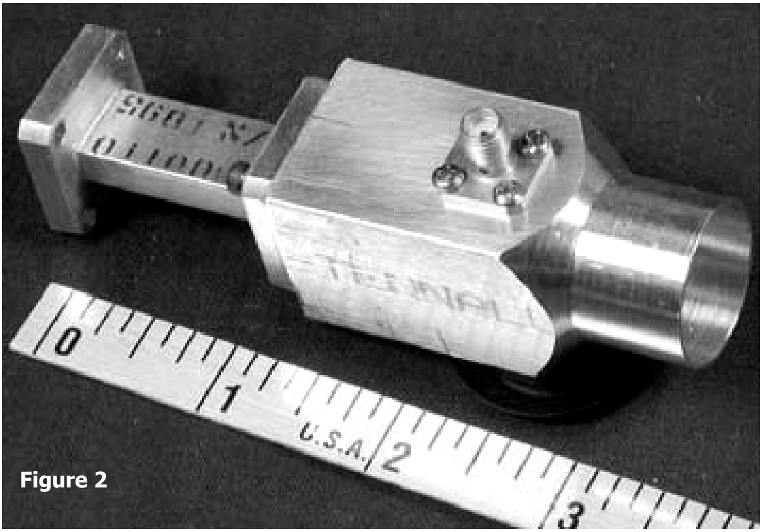


Figure 2

boring bar to minimize the shoulder so that I could at least measure the VSWR and make sure I was on the right track. Some improvement was necessary to make a proper taper. Matt, KB1VC, attempted to make a custom cutting tool, but the results were not encouraging – the countersink is clearly the right tool. I went back to the MSC catalogue and found a 7/8″ countersink, slightly oversized. The HSS tool steel is too hard to cut with ordinary tooling, so I used a tool post grinder to grind the countersink to the exact 20.4mm diameter.
Now we are able to machine the correct taper cleanly.

The basic machining procedure is:

1. Cut a piece of 1″ square aluminium to about 2.6″ long.
2. True up in 4-jaw chuck and face ends.
3. Drill a hole about 1/4″ diameter down the centre, all the way through.
4. Drill 3/4″ diameter about 35 mm deep.
5. Bore out to 20.4 mm diameter. This leaves a small shoulder at the taper.
6. Countersink with 20.4 mm, 60º countersink to form the taper.
7. Turn down outside diameter to leave about 1 mm wall, about 16 mm long.
8. Trim end to leave output section length of 35.4mm.
9. Drill out circular waveguide diameter to 9.2mm diameter.
10. Mark out WR-42 holes in back end, drill and tap 4-40 thread.
11. Mark out and drill SMA holes in one side, with centre 29.3 mm from open end.
12. Tap mounting holes for SMA 2-56 thread.
13. Clean up and degrease.
14. Fit SMA.

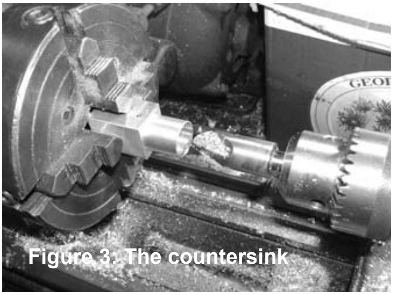


Figure 3: The countersink

Corrugated horns

The four corrugated horns I have ready to try are shown here in **Figure 4**.

I simulated the feedhorn with each of these, as well as with a plain conical horn. Patterns and calculated efficiencies are shown in the following figures at 24GHz, and at 10GHz with the phase centre at the best location for 24GHz.

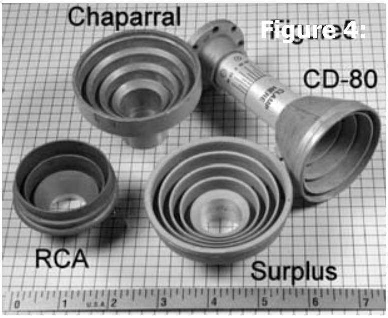


Figure 4:

The calculated efficiencies and phase centres are summarized in the table below**. (Fig 5)** At 10GHz, efficiencies are listed at the 24GHz P.C. (phase centre) as well as the optimum, assuming that the feed position would be more critical at 24GHz. Phase centres are measured from the centre of the aperture; negative numbers are inside the horn. All of the horns improve the performance at 10GHz while maintaining high efficiency at 24GHz. However, the best f/D and the phase centres for the two bands are not the same. The best choice appears to be the Chaparral horn, with phase centre differing by only 1.3 mm, so that the optimum position for 24 GHz is only 0.04λ off at 10GHz. The result is excellent calculated efficiency for both bands, 76% at 24GHz and 74% at 10.368GHz. This is comparable to the best single band feeds – a dual-band feed that does not compromise performance.

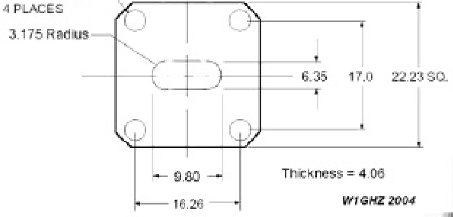
To make machining easier, a radius was added at the corners, and increased until the ends are simply 1/4 inch diameter; a slight increase in the wide dimension was required for the same impedance. The matching section improves the measured VSWR to about 1.05.

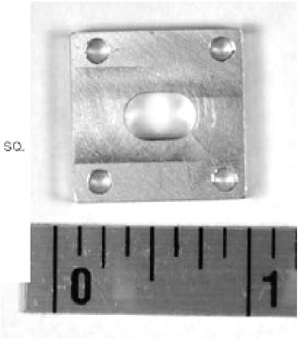
Figure 5:Dual band 10&24 GHz Feed - Calculated Horn Performance

	24GHz			10GHz			Offset f/D=0.7 at 24GHz P.C.	
HORN	**Effciency**	**f/D**	**Phase Centre**	**Efficiency**	**f/D**	**Phase Centre**	**24 GHz**	**10 GHz**
none	76%	0.72	1.2 mm	71%	0.38	0 mm	76%	47%
Conical	76%	0.8	-26	67%	0.7	-31.2	74%	66%
RCA	75%	0.93	-24.8	77%	0.7	-5.8	69%	68%
Chaparral	78%	0.81	-23.6	75%	0.65	-24.9	76%	74%
CD-80	77%	0.88	-29.1	78%	0.76	-4.9	68%	65%
Surplus	79%	0.8	-6.2	76%	0.7	-22	77%	70%

WR-42 Matching Section

Figure 6





I believe European 24GHz operation is at 24.048GHz; VSWR there is about 1.04. Adding the various corrugated horns to the basic feed has only a small effect on the VSWR on either band.

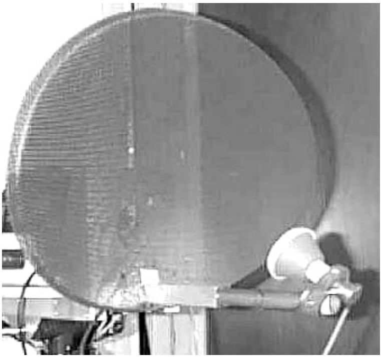
References:

1. Gary Lauterbach, AD6FP, and Lars Karlsson, AA6IW, "Dual-Band 10/24 GHz Feedhorns
for Shallow Dishes," Proceedings of Microwave Update 2001, ARRL, 2001, pp. 181-190.
2. Al Ward, W5LUA, "Dual Band Feedhorns for 2304/3456 MHz and 5760/10368 MHz,"
Proceedings of Microwave Update 1997, ARRL, 1997, pp. 158-163.
3. Joel Harrison, W5ZN, "W5ZN Dual Band 10 GHz / 24 GHz Feedhorn," Proceedings of Microwave Update 1998, ARRL, 1998, pp. 189-190.
4. Joel Harrison, W5ZN, "Further Evaluation of the W5LUA & W5ZN Dual Band Feeds,"
5. Proceedings of Microwave Update 1999, ARRL, 1999, pp. 66-73.
R.H. Turrin, (W2IMU), "Dual Mode Small-Aperture Antennas," IEEE Transactions on Antennas and Propagation, AP-15, March 1967, pp. 307-308. (reprinted in A.W. Love, Electromagnetic Horn Antennas, IEEE, 1976, pp. 214-215.)
6. www.mscdirect.com
7. www.ansoft.com
8. www.w1ghz.org

EDITORS NOTE: Paul's article has been somewhat shortened for BackScatter purposes. The complete article, including computer analysis charts of the horns, can be downloaded from Paul's excellent website as listed in reference 8 above.

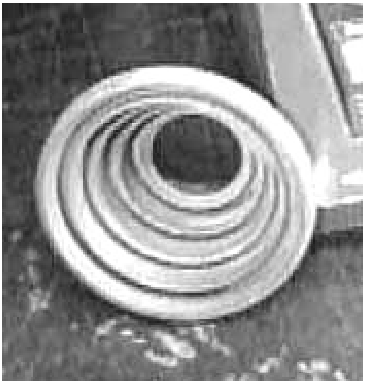
Using Surplus “Digi Dishes” at 10GHz

by John McCarthy, G7JTT



So far, using the G4JNT beacon, as a source, I get about an S1 signal with a 18" penny feed dish and a S5 with the digi dish shown in the photo on the left. I tried it with a G3PHO type dual mode feed but only got about an S1, the same as the penny feed. I then tried the 22mm open feed and got about S3 then tried various LNB feed horns and found one that got me an S5.

The feed is off an old, unbranded LNB . It is 63mm diameter at the open end down to 22mm copper at the other end.



There are 4 rings inside, with an average depth of about 6mm. The length is about 37mm and the copper feed enters about 5mm into the horn to give an SWR of about 1.4:1. As to the position of the feed, it is 330mm from the top of the dish to the centre of the horn’s face and 230mm from the bottom. It is pointed to the centre of the dish.

The gain seems to be in the order of about 4 "S" points above an 18" penny feed dish and about 5 "S" points above a 20db horn. It’s also much sharper than either the penny feed or horn, about 36 degrees from no signal to no signal. There seems to be at least two types of dish, I have one type (Channel Master) and my friend Noel, G8GTX, has the other (made by Grundig), so, over the next few months, we will see how they differ As to shape, the dishes differ quite a bit … mine had the LNB with a swept back elliptical horn where as Noel’s is a straight forward round feed.

At the Adastral Park (Martlesham) Microwave Round Table meeting this month, the following measurements were made on the antenna test range. The dishes appear to outperform the ubiquitous PW type with ‘penny’ feed.

CALLSIGN	DISH TYPE	GAIN (dBi)	COMMENTS
G7JTT	Channelmaster mini-dish with tubular horn feed	30.3	One of the popular digital Sky dishes. About 42cm wide
G8GTZ	Grundig Sky mini-dish with G3PHO/ W2IMU feed	30.3	Also one of the popular digital Sky dishes. About 42cm wide

Other Antennas

Designing Log Periodic Antennas

Kent Britain, WA5VJB

Over the last few years, I have developed 4 Log Periodic antennas from 50MHz to 6GHz. I wish I could say I have a secret design process, but there were many variations and metal tape on the early models. And while there are software programs such as IE3D than can analyze a log periodic antenna on a dielectric surface, I do not know of any programs that can design such an antenna. And as commercial antenna go, this is a poorly performing antenna.

Instead of a flat -10 dB Return Loss, these antennas have dips, at 2.3, 3.4, and 5.7 GHz. As a radio amateur, I am willing to loose a little performance outside the call frequencies for better spot performance on the call frequency.

If you would like a copy of the PCB file in Tango format, just send me an email at WA5VJB@FLASH.NET I'm sorry, but I only have the PCB file in Tango.

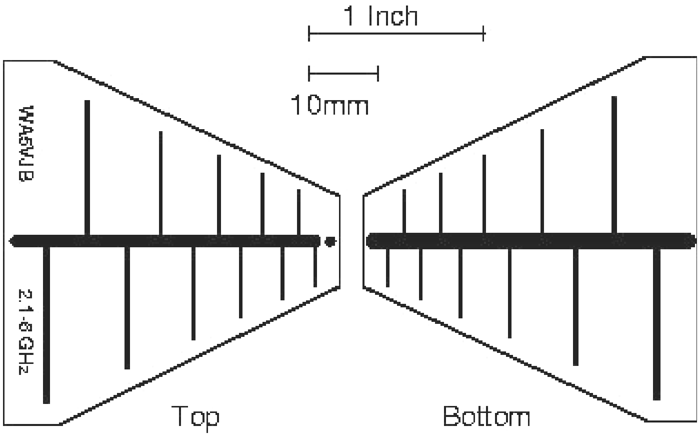
And now for a few of the problems designing PCB Log Periodics.

PCB Asymmetry

A Copper trace resting on a dielectric has a velocity factor. That is, a 1/2 wave dipole is shorter resting on Fiberglas than it would be in free air. While the textbooks give you the velocity factor for a stripline over a ground plane, this is not over a ground plane! But it turned out to be close to the velocity factor of a normal stripline, about 60%. However that is in just one direction. While the elements are about half their expected length, the element to element spacing is only reduced about 20%. So you cannot simply scale a free space Log Periodic by the PCB velocity factor and expect it to work well. As they say, "Been There, Done That"

PCB Dielectric Constant

I also learned the hard way that the dielectric constant of most PCB materials vary with frequency. So while FR-4 type materials have a dielectric constant (Er) of 4.2 to 4.4, it can drop to 3.9 at 2GHz, and even lower at



higher frequencies As you move up in frequency, the relative thickness of the material increased. At 900 MHz the PCB thickness (1.6 mm) is about .5% the length of the element. By the time I was up to 6GHz, the PCB material was up to 8% of the element length. So the Fiberglas dielectric has an ever increasing effect on the lengths of the elements, and companies just don't publish the Er of their non-Teflon PCB material above 1 GHz.

Booms

The velocity factor of a stripline varies with its impedance. That is, the wave travels at different speeds for wide lines and narrow lines. When the transmission line is soldered to the antenna, that side of the antenna effectively has less inductance and a higher velocity factor. You really don't want the waves travelling at different velocities on the opposite sides of the antenna! So the line trace on the transmission line side has been narrowed to allow for the effects of soldering on the coax.

Coax

The 900-2600 MHz version works well with either .141" semi-rigid or the smaller .085" coax. But the 2.1 - 6GHz versions get unbalanced with a really thick coax going down the boom. I recommend using either .085" semi-rigid coax on the 2.1 - 6GHz antenna or one of the Teflon RG-174 style coax with a diameter of less than .1 inch, or 2.5mm. The coax only needs to be a millimetre longer than the antenna, then you can transition to 7/8” Heliax? if you like. Keep the free tip of the coax as short as possible, and the shield as close to the via as practical. This improves phasing, especially at 6cm.

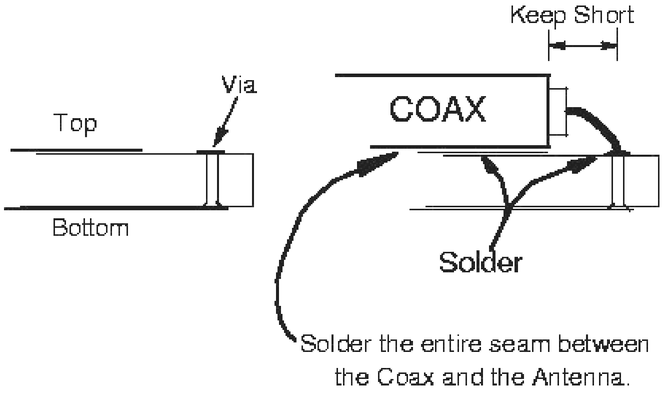
Grounding via at the back of the Log Periodic

Designing a Log Periodic to be grounded at the back of the boom helps solve many problems.

The entire antenna is at DC ground and many low frequency response problems are eliminated. In high EMI areas, ungrounded LP's can allow many VHF signals to become intermodulation sources. Removing the via will lower the lowest usable frequency slightly, but introduce many new resonance’s into the antenna. Removing the back grounding via is not recommended.

Power

I wish I could test one of these antennas to destruction, but I just do not have enough power. At 900MHz, the antenna has a relatively large active



area and could probably take 100 watts for short periods of time. As you go up in frequency, the active area gets smaller and smaller, and the dielectric loss goes up. At 6cm, the max power is probably in the 10 to 20 watt range. I am expecting some feedback from local EMEers.

Phase Centre

The phase centre is the point on the antenna that does the radiating. On an LP style antenna, the phase centre moves along the antenna as the frequency changes. Counting Elements from the small end, the 6 cm phase centre is about element #3. 9cm is about element #6, and 13cm is about element #9. One effect of the dielectric is the compression effect on the antenna, making it physically smaller. While the antenna has less capture area, the phase centre does not move as much as it would with a Teflon or air dielectric design.

Why didn't you build them on Teflon?

The short answer is money. If you know a PCB house that can make double sided PCBs with plated through holes from CAD files for less than $500 a panel, let me know! Also, I cannot use the current artwork. The elements will get about 20% longer, the element to element spacing will increase about 10%, and the asymmetrical booms would change. Both the 900-2600MHz and 2.1 - 6GHz versions went though the PCB house 3 times before I was happy, and a Teflon version would probably make several trips too. So a Teflon version would easily cost $2000 for me to develop.

As a Dish Feed

A Log Periodic is a compromise dish feed. How far down it is from an optimised feed depends on many factors, f/d, frequency, etc. But in general, a 3ft dish with an LP feed has about as much gain as a 2.5ft. dish with an optimised feed, and during contests you only have to carry one antenna. Mount the LP with the phase centre of the highest frequency you plan to use at the focus of the dish. The lower frequencies are slightly beyond the best focus point, but the loss is usually less than a dB. Also the dielectric compresses the antenna, so the phase centre does not move as much as it would with Teflon or an air dielectric Log Periodic.

The Flyswatter Microwave Antenna

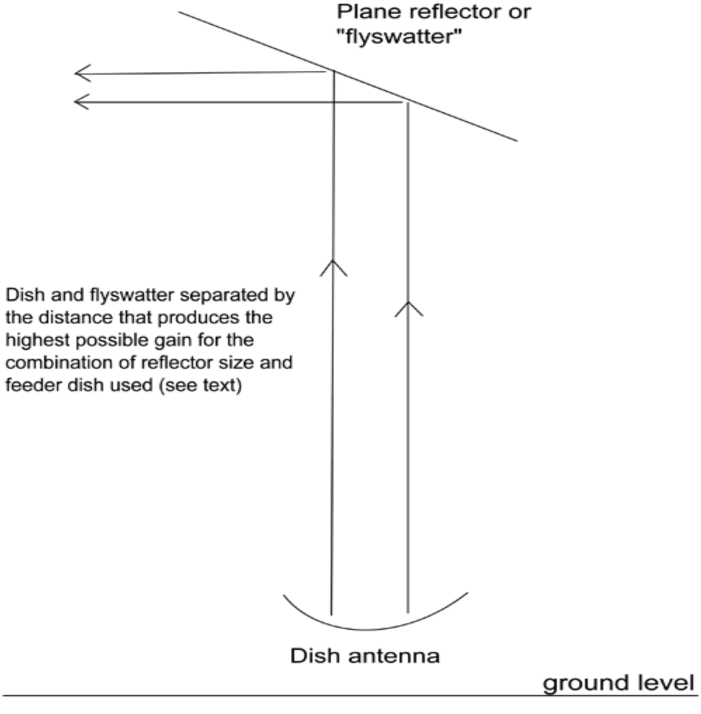
Peter Day, G3PHO

including notes from Doug, GW3ATM

In an effort to develop a decent 10GHz home station antenna system, Doug, GW3ATM, has turned to the Flyswatter Array. Pioneered many years ago in the UK by Mike Walters, G3JVL and Dain Evans, G3RPE, it does not appear to have received the attention it deserves. When correctly set up, the system offers a very good performance, often comparable to a moderately sized dish antenna at the same height.

Many of us have problems erecting a dish at a reasonable height because the most efficient way of feeding it usually involves having the microwave transverter either immediately behind or just below the dish and effectively out of reach for easy access.

The associated weatherproofing problems deter many folk from using this approach. Feeding the antenna via a long length of waveguide is another alternative but this usually means having some sort of rotary waveguide joint at the base of the waveguide and using a length of flexible waveguide or coax section to get around the rotator.


Plane reflector or
"flyswatter"
Dish and flyswatter separated by the distance that produces the highest possible gain for the combination of reflector size and feeder dish used (see text)
Dish antenna
ground level

Losses are necessarily greater when this latter method is employed.

The flyswatter antenna uses a very low loss feeder … air!

Basically, the microwave transverter is placed at or near ground level, where it can easily be adjusted and even taken indoors during periods of non activity.

Some distance directly above the transverter is a plane reflector, angled so that the radiation from the transverter's antenna (which is conveniently a small dish or even a horn) is reflected out horizontally to the surrounding area (see diagram right). The lower antenna doesn't need to be exactly underneath the plane reflector as the angles involved can easily be adjusted to compensate for any deviation from the vertical.

Small offset fed dishes provide excellent feed antennas for flyswatter systems since the transverter and its feedhorn can be mounted so that they do not point up into the sky from where they might get an input of rain during bad weather. They also tend to be more efficient than rear fed dishes using, for example, the "Penny" feed. The distance between the dish and the plane reflector is effectively a compromise based upon the relative gains of both dish and reflector (which in turn are a function of their physical dimensions). Unfortunately there is not room here to discuss this aspect in the detail it deserves! Paul Wade, W1GHZ, presented a fascinating paper on the subject at Microwave Update , held in Philadelphia, USA in September 2000. The paper can be found in the Proceedings of Microwave Update 2000 (available from the ARRL) and on Paul's outstanding website at: **www.w1ghz.org** . Interested readers should refer to either of these resources.

GW3ATM's flyswatter system consists of a 48cm diameter, ex-BSB, offset fed dish, mounted just above gutter level on a rotary, guyed pole, next to his bungalow. The dish is fed with an ex-Amstrad, sat-TV feedhorn (widely available at rallies). The aluminium, plain reflector is 1mm thick and measures 23 inches by 33 inches. It is fixed 11 feet above the BSB dish. A Sat-TV screw jack, fixed on a bracket on the opposite side of the mast to the reflector, is used to vary the tilt angle of the flyswatter. The pole, dish, plain reflector and transverter are so arranged that they all rotate together. If the dish and reflector did not move together there would be a change in the polarisation of the reflected signal. This problem was clearly demonstrated in Paul Wade's Microwave Update talk, when he used a laser pointer and plain reflector to "beam" the light at the four walls of the lecture room!

GW3ATM places his 10GHz transverter on a small platform, mounted on the same pole and just below the dish, so that a short and easy connection can be made to the offset feedhorn. Results so far have all been in receive mode only but have been very encouraging.

From his home at Llandogo, Monmouthshire, he can hear the Cleeve Common 10GHz beacon, GB3CCX, at good strength from two directions, via reflections from nearby commercial towers. He cannot hear this directly at ground level. G4UVZ has also been heard and he believes he can just detect the new Dorset beacon, GB3SCX. Doug is looking forward to the coming microwave "season" when he hopes to work many stations several hundred kilometres away. Rain scatter propagation should also provide very interesting results now that the reflector's tilt angle can be remotely adjusted.

The Flyswatter antenna system offers the home station operator a convenient and efficient way of making the most of his or her location.

Antenna Measurements at Adastral Park 2001

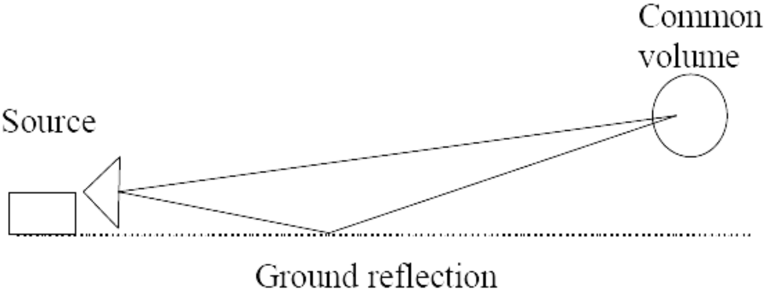
A report by Sam Jewell, G4DDK

Introduction

Having discussed at RSGB Microwave Committee meetings for several years the possibility of setting up an antenna gain measurement facility I finally ran out of excuses as to why it couldn't be done. So at Adastral Park 2001 I finally got round to setting up a 10GHz antenna range with the help of Kent, WA5VJB, and the loan of some critical items from Simon, G3LQR. I had decided early on that it would have to be limited to just one or two bands because of the sheer amount of equipment needed for each individual band. In this case it would be 10GHz and, time and facilities permitting, maybe also 2.4GHz. By limiting the measurements to one or two bands initially I thought I would be able to manage the logistics better.

The principle of short range antenna measurements has been known for years and antenna gain measurement on all bands from 50MHz to 24GHz has been a regular feature of VHF and Microwave events in the USA for many years. The results of these measurements are often reported in various ARRL publications and specialist newsletters. The principles have also been explained in many of these same publications as well as in a notable article by WA5VJB here in Scatterpoint, issue 1. Instead of trying to fight ground reflections, the normal bane of antenna measurements, the ground reflection is used in the measurement.

When the source is set up close to the ground then a ground reflection will occur, as normal. The direct and reflected rays combine in phase at some distance from the source and an increase in signal level will be observed. The distance can be calculated and then verified by probing this volume of space with the reference antenna. The volume is not too difficult to find and is shown by a volume of approximately 1 metre sides where the signal level does not change by more than about 1dB. In practice it is usually possible to find a common volume where the change is much less than 1 dB. Obviously, the volume



changes with frequency but not enough to be a problem to our measurement set up. Figure 1 shows the ground reflection range with the source antenna close to the ground and the common volume in which the measurements are taken.

The source is set up on 10368MHz and set to an output level somewhere between 10 and 100mW. The signal in the common volume, which is found by probing with the reference antenna, is indicated on the measuring 'receiver'. This level is noted and represents the reference antenna gain. In the case of the 10GHz measurements, the reference antenna was a 16.5dBi gain horn. A horn is used because the gain can be accurately determined from the physical aperture, assuming the flare angle is reasonable etc. The gain of my reference antenna is known to be 16.5dBi at 10.4GHz, from the data sheet anyway!

By substituting the reference antenna with the antenna to be measured (taking care to ensure the feed point of the unknown gain antenna is in the same place as the reference antenna was located when the reference measurement was made) the increase or decrease in signal level can be read off the meter on the 'receiver'. The gain of the unknown antenna is then the reference antenna gain plus or minus the difference on the meter in dB. Simple!

The problems

As you might expect there are some catches….

A power meter such as the venerable HP 432 could be used to measure the level of the signal received from the source. And, indeed, often is used. With this arrangement a pure CW source is all that would be required as the transmitter. However, the drift free dynamic range of the average power meter is very poor and the usual virtue of broadband operation is a serious problem.

With this simple (but workable) arrangement any stray RF will be indicated on the power meter with no indication of whether it is the wanted signal from our test source or unwanted pick up of the local 10GHz beacon, or even someone on a GSM phone, 2m handheld or the local taxi. This can lead to some very peculiar results if care is not taken to check the readings!

These limitations can be overcome by using a 1kHz (the standard) amplitude modulated source as the transmitter and a diode detector connected to a VSWR indicator as the receiver. All this is explained far better than I can do in issue 1 of Scatterpoint so I won't take up space by repeating it all, again. Suffice to say the modulated signal is not easily mistaken by the receiver and the use of a detected audio signal allows relatively drift free, highly repeatable measurements with a surprising dynamic range of nearly 50dB under optimum conditions.

For the 10GHz range I used a borrowed Marconi 6058B tuneable Gunn source. These were originally made as sources for driving slotted line measuring systems. As such they already have facilities for internal 1kHz amplitude modulation. The receiver was also a borrowed Marconi instrument. I used a Marconi 6593A VSWR indicator. These narrow band 1Khz tuned amplifiers with log detector are (subjectively) better than the better known HP415E VSWR indicators. I can confirm they are better as I had to give mine up to a well know American who insisted he needed it for measurements in the USA! Hence I had to borrow one to replace it. The second catch is that the accuracy of the range can depend on how close the gain of the unknown antenna is to that of the reference antenna. As the gain difference increases, so the possibility of errors increases. There are several reasons for this. A higher gain antenna is usually physically larger than our reference horn. This means it may not accurately 'capture' all the common volume because part of the antenna may be outside the common volume. However, this isn't too much of a problem because the contribution from the outer areas of the dish to overall gain is usually decreasing due to lower edge illumination. Also the bigger antenna should ideally be further from the source than our small reference horn. What this means in practice is that a 10dB gain difference is OK, 15dB is getting to be a problem and 20dB is asking for trouble. You will note that I asked in the original call for antennas for people not to bring 4 foot dishes to measure as these would (if working properly) be more than 20dB above the reference gain antenna of 16.5dBi. These larger antennas would probably show lower gain than actually achieved due to this effect of the range gain compression described above.

The results

So, now you know how the range works on 10GHz, the results. Please note, that these have already been published in both the RSGB Microwave Newsletter and the North Texas Microwave Society newsletter. In one of these fine publications the table was incorrectly formatted, prior to publication, and some information was lost. Hopefully, this is the definitive version.

2.4GHz

So, what about 2.4GHz and why 2.4? The first range I set up was at the University of Surrey, Amsat Colloquium 2001, I was assisted with the range by Dave, G0MRF. Dave arranged for a reference horn antenna to be made specially for the measurements. The antennas measured were all for use with AO-40 satellite and featured circular polarisation. The measurements went well and we had many converts to the technique of the measurements and I'm sure the range will re-appear at next year's Colloquium by public demand.

Probably most important, several very important lessons were learnt about helix antennas. How not to build them featured strongly!

The measurements were all published on the Amsat UK web site. Unfortunately, the reference horn antenna was not available for use at Martlesham Adastral Park and so a patch antenna was substituted. Although we knew the gain of the patch quite accurately, the broad beamwidth of the patch together with its low gain caused some problems and I am not too confident of the results obtained. For this reason I won't re-publish the figures in Scatterpoint although they have appeared in the previously mentioned publications.

Analysis of the results

The results are quite interesting.
I am confident that up to about 30dBi, the gains measured are quite believable.

Above about 30dBi the results look to me like there is some slight (~0.5dB) gain compression. I base this on the expected gain of the Andrew dish which should have been closer to 34dBi than the 33.1dBi measured. The loss of the coax to waveguide adaptor accounts for the missing fraction of a dB.

The PW dish and penny feed surprised me with how good it could be after years of believing that this arrangement was less than optimum.
?? I am impressed by the efficiency of the dual mode feed on the Amstrad 60cm dish. I'm definitely going to try this arrangement myself.

10GHz			
Callsign	*Antenna measured*	*Results (dBi)*	*Comments*
G3PHO	60cm Amstrad dish with W2IMU feed and extension	35.2	G3PHO version
G3PHO	As above without extension	34.2	
G8PSF	80cm Amstrad dish with W5LUA dual band feed	30.5	10 + 5.7GHz feed
G3LYP	BSB 35cm dish with feed based on LNB feedhorn	29.5	
G3LYP	PW dish with penny feed	29.5	After adjustment
G3LYP	Horn	19.8	20 dB commercial design
G8LSD	80cm prime focus with penny feed	34.5	
G3JMB	60cm prime focus with penny feed	27.5	
G3JMB	60cm prime focus with dipole and reflector	27.5	
G3JMB	PW dish with penny feed	29.0	
G3NYK	Horn	21.5	Commercial horn with large aperture
G3NYK	Elliptical Horn with Fresnel lens	19.5	Commercial
G3NYK	Small horn	15.8	Commercial
G4DDK	Andrew 60cm prime focus with shepherd crook and horn feed	33.1	The popular commercial surplus dish
G4DDK	RSGB '20dBi' horn from the Microwave manual	20.0	Calculates at 19.9dB at 10368MHz
G4ZXO	42cm dish with dipole and reflector feed	30.0	
G7JTT	Channelmaster minidish with tubular horn feed	30.3	One of the popular digital Sky dishes. About 42cm wide
G8GTZ	Grundig Sky minidish with G3PHO/W2IMU feed	30.3 !	Also one of the popular digital Sky dishes. About 42cm wide
G7JTT	16 slot (8 each side) waveguide	12.3	Typical beacon antenna
G3XGK	Horn	19.0	Commercial
G3XGK	25 inch prime focus with penny feed	32.0	Home brew fibreglass design

Lessons learnt

What lessons were learnt from the 10GHz range at Adastral Park?

The most important was to have a go. We have all assumed for some time that you just can't do accurate antenna gain measurements on an amateur test range. This assumption has persisted too long. Let's hope it's been put to bed once and for all.

There is great satisfaction is seeing the gain of your precious dish/horn/WHY confirmed by measurement. I would guess that many of the onlookers started thinking about trying that dual mode horn after all. It might just make it possible to work that extra path that doesn't quite go now.

Even small dishes such as the Sky minidish give very acceptable gain when correctly fed. Just the job for that mast head system where there isn't quite room for a 90cm dish on the short and already overloaded stub mast.

Chapter 4

Low Noise Amplifiers

Fighting for the Last Few Tenths of s dB in LNA Noise Figure .. Is it Worth it?

Richard T, Nadle, K2RIW

***Editor's note:** Dick posted this interesting discussion document on the Internet. We thought it so thought provoking that we just had to publish it in the Newsletter. While it is aimed primarily at the EME operator, there is much for all of us to consider. Our grateful thanks go to Dick for permission to use his document.*

Introduction

Some of this information came from the "Do You Need a LNA Line Driver" presentation by K1FO at the 3/17/01 NEWS (North East Weak Signal) Meeting at Enfield, CT, USA. In many amplifying devices, there is a considerable difference between tuning for the best gain, and the best Noise Figure (NF). Under certain conditions it is possible to tune a particular amplifier for best gain and realize a NF of 0.9dB. That same amplifier may deliver 0.3 dB NF, with slightly less gain (maybe 0.5 dB less), when tuned for best NF. At first you might say, gee, a 0.6dB NF improvement! Is that small improvement worth fighting for? The answer is YES, YES!

In the paragraphs that follow I'll attempt to show you:

(1) Why it is worth fighting for.
(2) Why it is hard to detect the improvement with typical laboratory equipment.
(3) What else in the system must be working right to get the full benefit.
(4) The Antenna Impedance problem.
(5) You'd better have the correct Gain Distribution.

(1) Why is it Worth Fighting For

You're about to have a demonstration that dBs of NF are not linear, when dealing with a cryogenic communication system and a small NF improvement can make a big difference in the Signal to Noise Ratio (SNR). A communication system becomes cryogenic when you aim the antenna into cold space.

Two EME System Examples -- Assume I have an excellent EME antenna that is presently aimed at a high elevation angle and it is aimed at a cold direction in the Universe, where the true Celestial Background is 2.73 Kelvin (the residual of the Big Bang, by Penzias and Wilson, Nobel Laureates, 1968). Under those conditions, that excellent EME antenna might have a total Antenna Noise Temperature (Ta) of 30 Kelvin (they [the experts] call it 30 Kelvin).

System (A) -- If your Low Noise Amplifier (LNA) had a NF of 0.9dB, that equals an electronic Noise Temperature (NT) of 66.78 Kelvin. The formula for NT is:

NT = [ALOG(NF/10) - 1] * 290.
Your total possible system NT (Ts) is now the sum of the Ta and the NT.
Ts = Ta + NT. Ts = 30 + 66.78 = 96.78 Kelvin.

System (B) -- If I were using that LNA when it is tuned at a NF of .3dB, the LNA's NT would equal 20.74 Kelvin. Now the Ts = 30 + 20.74 = 50.74 Kelvin. The only thing that is always linear about a cryogenic communication system is that the total Noise Power is proportional to the Ts in Kelvin. You have to almost ignore the dBs.

When I compare a system with a Ts of 96.78K (the A system) to a system with a Ts of 50.74K (the B system), that ratio is 2.80 dB. In other words, that EME signal will be 2.80dB further out of the noise when I use the LNA with the better tuning (I'm ignoring Moon Noise). That 2.80dB of difference can easily make the difference between a QSO and a missed QSO, when I'm listening to a weak EME signal.

A Better EME Antenna -- If I were able to make further improvements to that EME antenna so that the Antenna Temperature (Ta) was 20 Kelvin (and that is possible), then that ratio would be 86.78 Kelvin versus 40.74 Kelvin, or 3.28dB of system improvement in Signal to Noise Ratio (SNR). All of this is from a 0.6dB NF improvement (0.9dB to 0. dB NF) in the LNA! A ~ 3 dB SNR improvement for a 0.6dB NF change, that's a beautiful (apparent) non-linearity but it really isn't non-linear … it just looks that way (in dBs).

(2) Why is it Hard to Detect With Typical Lab Equipment -- When I make measurements with room temperature laboratory equipment, everything (the pads, signal generators, FM Receiver, etc.) is at approximately 290 Kelvin. In that 290K environment, a change in the LNA's NF from 0. dB to 0.3dB, can create a change in SNR of 0.6dB (at best). Here the dB' are linear. Unless you are using some good laboratory equipment (such as a SINAD Meter or a good Automatic NF Meter) you will probably not detect that 0.6dB SNR improvement. Your ear probably doesn't have enough discrimination to allow you to hear it, when you hit that "Sweet Spot" in NF tuning. As section (3), (4) and (5) will show, you will not be able to realise the benefit of that 0.6dB of NF improvement, unless the rest of the receiver has a NF of nearly 1.0dB or so, has the proper antenna impedance and the proper gain distribution.

(3) What Else in the System Must Be Working Properly to Get the Full benefit -- If I were using a receiver that was Gain Starved or had a second stage NF of 15dB, then the tuning of the LNA takes on an entirely different characteristic.

High Transceiver NF -- Many of the currently used Base Station 2m and 70cm Transceivers have a bare foot NF of 12 to 15dB. This occurs because Japan favours dynamic range over receiver sensitivity. Those Transceivers are front end Gain Starved. If you lived in a dense community where there was a Ham Radio Operator living on each street, you might agree with this approach.

Add an LNA -- Therefore, almost every American SSB operator must add a ~ 20 dB gain LNA in front of his Transceiver, if he desires full sensitivity of his communication system. If he doesn't add that LNA, all his SSB friends will eventually call him an Alligator (he is all mouth), instead of a Rabbit (a guy that is all ears). During a terrestrial contest, everybody can hear that Alligator call CQ, and they answer him, but because of his poor hearing aid (receiver), he only hears the locals and only responds to them.

The rest of the contest operators become frustrated, and learn to ignore him. Unless some local explains this to the "Alligator," he will conclude that there wasn't much activity during the contest. It is well known that the total system's cascaded NF (NFs) is equal to:

NFs = NF1 + (NF2-1)/G1 + (NF3-1)/(G1*G2) + ...,
where:
NF1 = NF of the first stage (as a real, anti-LOGed number).
NF2 = NF of the second stage (as a real number), etc.
G1 = Gain of the first stage (as a real number).
G2 = Gain of the second stage (as a

real number), etc.

Sometimes Best Gain = Best NF - If I were tuning a communication system's LNA, while it is connected to that bare foot Transceiver, I would find that the best system NF would be approximately the LNA tuning with the maximum gain. Here is an example: If you experiment with just the first two terms of that Cascaded NFs formula, and use 15dB for NF2, 0.3dB for NF1 and a gain of 10dB for G1, you will find that the System Noise Figure (NFs) [in dBs] is equal to 6.16dB. If, now, I use 0.9dB for NF1 and 10.5dB for G1, I realise a NFs of 5.98dB. Notice, that I worsened the LNA's NF by 0.6dB, while improving the gain by only 0.5dB, yet the system NF IMPROVED by 0.18 dB.

What this example demonstrates is that when your receiver system is gain starved, and has a high second stage NF, then the LNA's gain is much more important than it's NF. Even if I were using perfect laboratory instrumentation (such as a perfect NF Meter) while tuning that LNA in that environment, I would end up tuning it for near maximum gain, not best NF.

The conclusion is that to get the maximum benefit of a Super Low Noise LNA, you have to put it into the right environment while tuning it, and using it. Otherwise, you may be "casting pearls upon swine" ... you could be wasting your time and your money.

(4) The Antenna Impedance Problem -- It is well known that when you are tuning an LNA for the best NF, you are primarily adjusting the impedance that the front end of the LNA is looking into.

It would be quite wasteful to carefully adjust an LNA stage while it is connected to perfect 50 ohm resistive laboratory equipment, and then connect it to an antenna with a VSWR of 1.41:1. That 1.41:1 antenna could be an impedance that consists of a capacitive (or inductive) reactance of 17.4 ohms in series with a resistance of 50 ohms, or it could be a pure resistance of 70.7 ohms, or 35.4 ohms. In any of these cases, that 1.41 VSWR antenna would drastically change the LNA's NF. Some of those possible impedances would cause more of a NF detriment than others, depending on the particular LNA design.

That uncorrected 1.41:1 VSWR antenna could easily raise your LNA's NF from 0.3dB to 0.9dB and hurt your EME receiver sensitivity by ~ 3dB. However, on transmit, the 1.41:1 would only cost you 0.127dB of transmission loss.

Your choices for correction are either to perfectly impedance match the EME antenna (with a double stub tuner, for instance) to make it look like a 50 ohm resistive load to the LNA, or do the NF tuning of the LNA while it is connected to the antenna -- such as by injecting the NF Meter's Noise Source through a 2 dB Directional Coupler (DC) that is always left in the antenna line.

The DC Line Perturbation -- If, after the NF tuning of the LNA, you made the mistake of removing the Directional Coupler, you would be changing the transmission line length, and that would rotate the antenna impedance to a different place on the Smith Chart. This would disturb the LNA’s NF tuning.

Another solution is to add a carefully chosen length of line to the DC's straight through path, so that the DC plus the extra line is an exact multiple of a half wavelength (electronically). Now you could remove that DC plus extra line, and not effect the antenna's impedance.

(5) You'd Better Have the Correct Gain Distribution -- To realise the system's best possible sensitivity requires that you have enough front end gain and a low enough second stage NF but this requires a compromise of system NF versus Dynamic range. You usually can't have both all at once. The best possible system NF usually requires a lot of front end gain (sometimes 20 to 30 dB).

But, a system with that much front end gain will saturate 20 or 30 dB sooner from strong local signals -- that's the problem that the Japanese equipment manufacturer's discovered.

Noise Power Saturation -- Also, bear in mind that even if you live in the "Out Back," and saturation from local operators isn't a problem, there can be another subtle detriment from the use of super high front end gains -- Noise Power Saturation. It is possible that the later stages of your receiver are being subjected to so much Noise Power, from all the front end gain, that they are beginning to saturate on the instantaneous noise peaks. Even if that saturation is only a fraction of a dB, it can lower the SNR of a weak signal. It is well known that a limiting stage will suppress a weak signal that's surrounded by noise with, what is called, "Signal-Cross-Noise Terms." In other words, it is possible for a super high gain system to suppress that weak EME signal you're trying to hear, in the later stages of your own receiver. This phenomenon is quite subtle, and not easy to detect. But, if the gain in your system is shoving the S Meter above S7 from the basic Noise Power, than be wary, it could be happening to you. The only quantitative test procedure I know of to detect this condition is the "Notched Noise Power Fill-In Test," also called the Noise Power Ratio (NPR) Test.

IF Filter BW -- It is also possible that your system is going into and out of Noise Power Saturation, as you change the bandwidth of the IF filter. At first, you would think that the broader IF filter selection would aggravate the problem. However, it is possible that the more narrow filter selection allows less noise power into the final detection stage, and this in turn causes a smaller AGC voltage, which increases the receiver's gain, and causes Noise Power Saturation in an earlier stage.

Sun Noise Problems -- As your EME system becomes more refined, and you experience a larger number of dBs of Sun Noise measurements, it is possible that with the added Sun Noise power, your receiver system could be experiencing Noise Power Saturation. That would give you a pessimistic Sun Noise measurement. One simple method of detecting this problem would be to put a 6dB pad in various places (after the LNA), and repeat the Sun Noise measurement. If you get a better reading, you may have the problem.

The best system for high dynamic range is one that has a gain distribution that's just enough, at each stage, to override the NF of the next stage. The best system NF requires considerably more front end gain than that. Soon we will all pay more attention to the Noise Power Saturation characteristics of our tuneable IF receivers. Then we will simultaneously have the best system NF and high dynamic range.

A state-of-the-art 2.3GHz Pre-amplifier

Grant Hodgson G8UBN

Eagle-eyed readers may have noticed the reference to an ATF-54143 in John G3XDY's notes on the pre-amplifier testing results at the November Adastral Park round table, published in the March 2002 Newsletter. Here are some more details of this pre-amp which has some rather interesting properties :-

The ATF-54143 is the first of a new breed of low noise GaAsFETs. Released by Agilent Semiconductors (formerly HP) in mid-2001, it is less than one year old and offers some remarkable properties. It was designed for the mobile phone base station market, where low noise and good strong signal handling ability have to be achieved at the same time. However, it can be used up to at least 6GHz (although the gain is starting to roll off at this point), and is ideally suited for amateur microwave use in the lower and middle bands.

Agilent call the device an Enhancement-mode Pseudomorphic High Electron Mobility Transistor, or E-PHEMT. HEMTs have been around for some time now, and will be familiar to anybody who has built a microwave low noise amplifier (LNA). 'Pseudomorphic' is a development of the basic HEMT, and refers to the way the Gallium Arsenide is doped during device manufacture; the channel of the FET being made from many thin layers, which form a lattice structure. It sounds very impressive but is only of relevance to the semiconductor physicists involved in the details of the design and fabrication of the device itself.
(Incidentally, PHEMT is pronounced 'pee-hemt - not 'femt, but there doesn't seem to be a universally accepted way to pronounce EPHEMT'!) The really interesting bit is 'Enhancement'. Until now, all low noise and high power GaAsFETs for
RF and Microwave use were of the 'depletion' type, which means that the device has to be biased into the operating region by ensuring that the gate is at a more negative potential than the source.

The two most usual ways of achieving this are by grounding the gate at DC and using resistors in the source lead(s), or by grounding the source lead(s) and using a separate negative voltage generator. Enhancement mode FETs require a positive voltage to be applied to the gate, which obviously makes things far easier in terms of circuit design and construction.
The second interesting feature of the ATF-54143 is it's strong signal handling performance.

Traditionally, low noise GaAsFETs have had significantly worse performance in terms of being able to
handle either in-band or out of band strong signals than bipolar transistors or MOSFETS. The parameter most often used to describe strong signal handling performance is the 3rd order intercept point IP_3, either referred to the input (IIP_3) or the output (OIP_3) [1],[2],[3].

Note that the difference between the IIP_3 and the OIP_3 is simply the gain or loss of the device, so that an amplifier (or transistor) with an IIP_3 of +10 dbm and 6dB gain will have an OIP_3 of +16dBm.

The ATF-54143 has an OIP_3 of up to 37dB at 2.3GHz, with an associated gain of 16dB, giving an IIP_3 of +21dBm which is a very impressive figure indeed. The IP_3 is a function of the bias conditions, and this allows a trade-off to be made between IP_3, gain and noise figure.

On the higher microwave bands, strong signal handling is not usually an issue but there are at least two cases where good strong signal handling could be used to good effect, particularly at 1.3GHz :-

1) The 1.3GHz band has a close proximity to the frequencies used by the huge Civil Aviation Authority radars in some parts of the country. Some 1.3GHz receivers are consequently overloaded.

2) Repeater builders have to go to great lengths to ensure that the repeater output does not de-sensitise the receiver. Traditionally repeaters have required a low loss, high Q cavity duplexer in order to separate the Tx and Rx signals. A receiver front end with a very high IP3 may not be as susceptible to self-desensitisation, allowing the possibility of a lower specification receive filter.

The ATF-54143 could help in both of these situations. The noise figure of the ATF-54143 depends on frequency, but at 1.3GHz the NFmin (minimum noise figure if the rest of the circuit was ideal) is only 0.4dB, and only 0.5dB at 2.3GHz. This is not quite as good as some other devices, such as the ATF-36077, but for the vast majority of cases this will not matter for two reasons :-

i) Noise figure differences of one or two tenths of a dB are only relevant in very special cases for space communications, such as satellite and EME - for terrestrial applications, it makes no real difference to the received signal/noise ratio due to the relatively high level of background noise (approx. 290K).

ii) Use of good RF circuit design techniques, high Q microwave passive components and low-loss PCB materials can reduce the circuit losses in a new LNA design such that the degradation in noise figure due to the these components (i.e. components external to the active device) are very small indeed, typically less than 0.2dB.

'So what's the catch?'

The ATF-54143 is not expensive, currently being only £6 each, which is considerably cheaper than (for example) the Mitsubishi MGF1402. One of the biggest problems is the size of the device - it is obviously in a surface mount package, as with all new devices (apart from those with no package at all - i.e. bare die!), and the package of the ATF-54143 is very small with the leads on a 0.65mm pitch, so some form of optical aid is required when soldering. There is also the usual problem with all HEMTs in that there is a reasonable amount of gain at high frequencies, and this gain increases as the frequency is reduced. This leads to the possibility of instability anywhere from several hundred MHz to over 10GHz, and careful circuit design is required to ensure that the resulting amplifier is stable.

Circuit Description

The positive gate bias can most easily be derived from a simple voltage divider consisting of two resistors. However, the drain current is highly dependent on gate voltage, and the relationship between the drain current and gate voltage (transconductance, Gm) varies from device to device, as do the individual I-V curves. The ATF-54143 data sheet [4] gives details of an active bias circuit, which ensures that the bias conditions (drain voltage and current) are consistent from device to device, and offers a degree of temperature stability. The design presented here is based on the Agilent design, with some subtle modifications.

Referring to the circuit diagram **(figure 1)**, R1 and R2 form a potential divider which keeps the base of TR1 at a constant voltage of approximately

2.7V. The emitter voltage of TR1 is simply the base voltage + 0.65V. This sets the drain voltage of TR2 at approximately 3.4V.

The drain current is set by the resistor R3 to be approximately 30mA. The gate current is almost zero, and can be ignored for the purposes of biasing. The gate voltage is therefore the same as the collector voltage of TR2, and is regulated by TR2 such that conditions for the drain voltage and current are always met. If the drain current of TR2 was to rise for any reason (such as a change in temperature), the voltage at the emitter of TR1 would drop causing the voltage at the collector to drop. This would reduce the gate voltage, causing the drain current to drop, thus maintaining bias stability.

However, the emitter-base voltage of a bipolar transistor decreases as the temperature increases.

Therefore, D1 is included to compensate for this; if the temperature increases, the voltage drop across D1 will reduce, causing the base voltage of TR1 to increase, but the emitter-base voltage of TR1 will also decrease with increased temperature, and consequently the voltage at the emitter of TR1 will be almost constant over a wide range of temperatures. Measurements show that with this arrangement, the drain voltage of TR2 varies by only 100mV and the current varies by only 1.8mA over the temperature range -18°C to +60°C. The temperature performance of microwave circuits is often ignored, but masthead mounted pre-amps can be subject to extremes of temperatures, especially if mounted at the feed point of a dish which is pointed at the sun in order to make G/T measurements. The active bias circuit ensures that consistent bias conditions will be maintained over a wide range of temperatures. Note that this bias circuit could be adapted for use with a negative gate-biased FET to give a similar level of bias stability, but the author is not aware of any amateur designs that actually do this.

The bias conditions of V_{DS}=3.4V and I_D=30mA were chosen to give a low noise figure and reasonably high gain at 2.3GHz, the objective being to reduce the noise figure of the system as a whole. The bias conditions are set only by R1, R2 and R3.

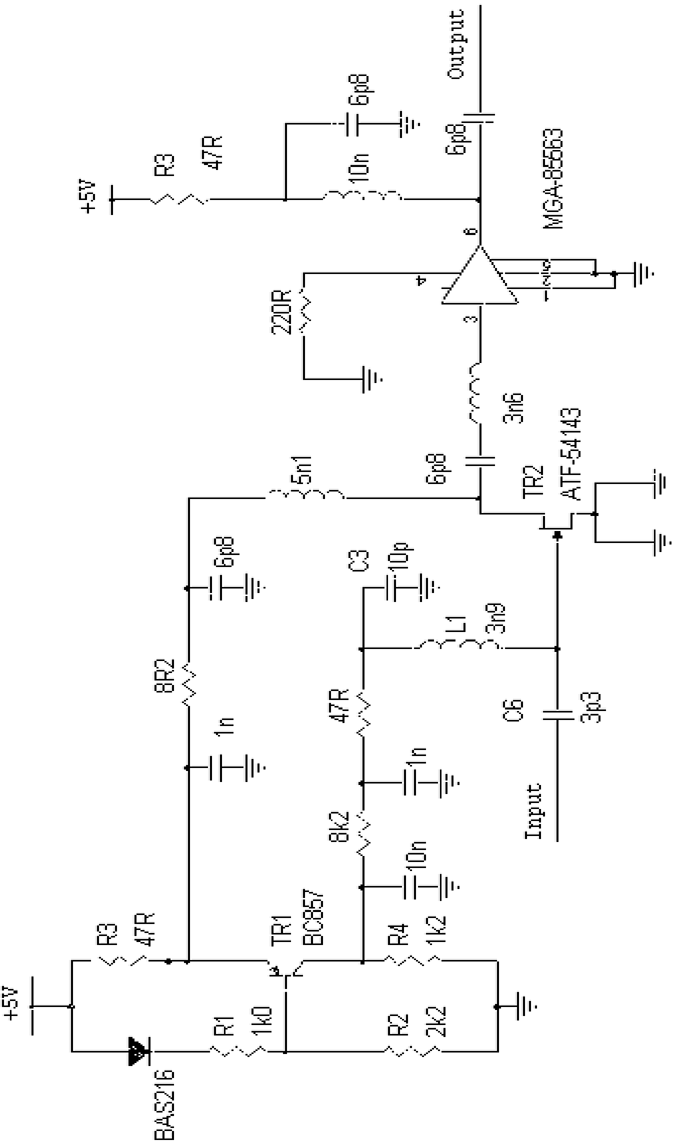
Source Degeneration

Traditionally, HEMTs and PHEMTs have required the absolute minimum inductance from the source lead(s) to RF ground for best performance. This is why FETs usually have two source leads; the inductance to ground is reduced which increases the gain. The ATF-54143 has been designed to allow some source inductance to be used in order to raise the input impedance to be closer to 50 ohms. The input impedance is of little importance in the majority of amateur radio applications, the exception once again being in repeaters where the performance of the duplexer is a function of the load presented to the ports.

For the majority of amateur applications, the input SWR of a low noise amplifier is of secondary importance to the noise figure and gain. However, the PCB (see below) has been designed to allow some source degeneration if required.

This is achieved by etching two parallel tracks for each of the source leads. One track connects to the device source lead, the other is connected to ground with a row of vias. A small (0402 size) zero ohm link is used to connect the source lead to ground; the position of the link setting the source inductance. For 2.3GHz (and above) the inductance is set to minimum, although the resulting impedance is not zero – the size of the zero ohm link acting like a short transmission line with a nonzero inductance. This causes a reduction in gain of approximately

Figure 1: 2.3GHz preamp: circuit diagram



2dB, although the noise figure is unaffected. More details of source degeneration can be found in reference [5].

Input match

As with all discrete low noise amplifiers, the optimum noise figure does not occur when the device is matched to 50 ohms. This causes all manner of confusion to some, but suffice to say that the best (lowest) noise figure is obtained when a certain mismatch occurs. The input impedance is transformed to this optimum impedance with C6 and L1; C3 ensures that the junction of L1 and R6 is at RF ground. In order to achieve the best noise figure possible, the losses associated with these input matching components must be kept to a minimum, which means that high Q components must be used. Traditionally, inductors at these frequencies would either be printed on the PCB or formed from wire of the correct size. The Q of printed inductors is limited by the loss tangent of the PCB material, and it is difficult to change the inductance of a track once the PCB has been manufactured. The use of surface mount wound inductors allows for very high Qs, and also allows for a single PCB to be used at different frequencies. The cost of wound inductors have recently been reduced to partly due to intense competition between the manufacturers, which is good news for the microwave constructor.

There are no variable capacitors or inductors that need to be adjusted for best performance – this is a true 'no-tune' design.

Recent advances in ceramics technology (developed for the mobile phone industry) have led to the development of surface mounted inductors in a range of different values and sizes, some of which have a very high Q at low, and even middle, microwave frequencies. Once again, the bad news is the size of these parts – the highest Q inductors are in the 0603 or 0402 size. Whilst these are 'industry standard' sizes, they do pose some problems for the amateur at home – 0603 components measure approximately 1.6 x 0.8mm; 0402 components are considerably smaller! Hand soldering is possible with the right equipment, but the techniques are very different to those required for more conventional, larger components.

The inductor chosen for the input match in this design is a Coilcraft 0603CS-030 which has an inductance of 3.9nH +/-5%. The actual inductance varies with frequency, but the graph of L vs. f is very flat and the actual inductance at 2.4GHz is very close to 3.9nH. The Q of this inductor at 2.4GHz is approximately 100, and this high Q value is a major factor in keeping the noise figure of the amplifier down.

Second Stage

Some applications will benefit from a second stage of amplification. This is achieved with an Agilent MGA-85563 low noise Monolithic Microwave Integrated Circuit (MMIC). The MMIC is fairly easy to use compared to a discrete FET, requiring only one inductor to match the input and an RF choke and some decoupling capacitors on the output. An external resistor sets the bias current, and that's it!

The bias current of the second stage has been set to approximately 15mA. The second stage has a noise figure of approximately 1.6dB and a gain of 18dB. The IIP_3 of the MGA-85563 is approximately –7dBm, which is considerably worse than the OIP_3 of the ATF-54143. This means that the overall system IP_3 is limited by the second stage, not by the first stage, and the resultant IIP_3 is approximately –20dBm. However, as mentioned previously, the strong signal handling performance at 2.3GHz or 2.4GHz is not really an issue, and this level of performance will be found to be more than good enough.

'Universal' Printed Circuit Board

The PCB used for this project has been manufactured professionally with 0.4mm plated through holes and a tin-lead finish. This is obviously not the cheapest solution, but it is felt that this was only option given the small size of some of the components used. Professionally etched PCBs offer a high degree of repeatability which is essential for consistent performance, and completely eliminate any uncertainty due to etching errors or mistakes.

The PCB dielectric is a material called AR320, made by Arlon Inc. in the USA. This is a hybrid material of PTFE and fibreglass, offering some properties of both. The losses are much lower than with conventional FR4, but the cost of the material is much less than even the most common PTFE only material, Rogers Duroid 5880. Also, the hybrid material is much stronger than Duroid. The dielectric constant is (perhaps not surprisingly) about half way between the two, at 3.2 (hence the name AR320) and the thickness is 0.8mm. The PCB uses surface mounted components exclusively.

This allows the same PCB to be used for different bands; all that is required is for use at different frequencies is to change the value of the matching components, and possibly the bias resistors. It is hoped that the same PCB can be used for a number of different pre-amps from 432MHz or below to 3.4GHz and maybe even 5.7GHz. It is fully appreciated that 0.8mm Arlon AR320 is an unusual choice of dielectric material. This material was chosen for it's suitability to do the job, and also because the author has stock of this material, it being used on other projects. It is certainly not as readily available as FR4 or Duroid, and so 'home-brewing' the PCB will not be easy, if it is possible at all. (There may be an alternative PCB material such as Rogers 4003, although this has not been tried). However, this LNA is presented to show the levels of performance that can be achieved with a state-of-the-art design, not necessarily something that can be etched in the kitchen sink. If any constructors wish to have a go at etching their own PCBs then please get in touch with the author who will supply artworks. It may be possible to modify the PCB artworks for different dielectric materials, but this has not been tried and there are many pitfalls for the unwary.

Performance

The noise figure of the ATF-54134 increases with frequency, and much work has been done to reduce the noise figure at 2.4GHz to an absolute minimum, for use with the AO40 S-band downlink at 2401MHz. The resultant noise figure of the two stage amplifier is 0.6dB (Te=48K) and the gain is approximately 28dB. The ATF-54143 can be used as a single stage amplifier, in which case the noise figure is slightly less, and the gain is 13dB. The output of the first stage (and the input of the second stage) is matched to 50 ohms, making it easy to use the first stage on it's own if required. The simulated IIP_3 of the two stage amplifier is -20Bm. This is yet to be measured.

Further Developments

A 1.3GHz version of the pre-amp is currently being developed. The noise figure is expected to be approximately 0.4dB, although the biasing will be changed in order to improve the IP_3 performance which may lead to a slight increase in noise figure. The second stage will be re-designed in order to improve the IP_3. There are other reasons why a high IP_3 is of advantage at 1.3GHz; full details will appear in a future edition of the newsletter.

Conclusion

The above design demonstrates that the amateur microwave community can benefit from the massive amount of research and development that is being carried out by the major component manufacturers. New devices are being introduced literally on a daily basis, and some of these devices can be put to good use by radio amateurs.

Other designs for amateur radio use of the ATF-54143 have been published [6], but the author is not aware of any other designs for the 2.3GHz band using this device.

The author does not claim any originality for this design; all the individual ideas and circuit blocks are used elsewhere in commercial receiver designs. However, these ideas are now put together and demonstrated in a practical way for the amateur microwave experimenter.

Apart from the very small size, the ATF-54143 has no real drawbacks. Now we have a device with very low noise figure, positive gate bias, superb strong signal handling and easy availability. Active bias circuitry ensures that bias adjustment potentiometers are not required, and the use of high Q lumped components gives a low noise figure and no need for tuning. So not only can we have our cake, we can eat it and have second portions as well!

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6. Dubus Q1/2002

6cm Wideband Amplifier Using a Surplus Sat TV LNB

Mike Parkin, G0JMI

The wideband amplifier described here is based on the two-stage GASFET amplifier taken from an Amstrad "Blue Cap" Low Noise Block (LNB) satellite TV down-converter.

Performance

Measured gain for 1mW drive at:
2320.1MHz: 17dB approx.
3400.1MHz: 17dB approx.
5760.1MHz: 15dB approx.
Typical output power out:
2320.1MHz: 60mW
5760.1MHz: 50mW
Measured using HP Power Meter 430c into 50 ohms load.

Summary of Construction

The GASFET amplifier was built using two of the three amplifier stages from an Amstrad "Blue Cap" LNB TV receiver that link the input to the mixer.
The LNB module was obtained from a Rally for about £1. The pcb is removed from the LNB's box with access to it being gained by drilling out the four outside rivets and then removing the screws from the internal screening.

The screw holding the +5 Volts d.c. regulator needs to be removed and the output wire cut before the pcb can be removed from the box.

Using a sharp knife or scissors cut out the two-stage GASFET circuit as shown in **Figure 1**.

Remove the copper track runs for the integral tuned circuits and track runs as shown in Figure 1.

A pcb drill fitted with a small grinding tool was used for this purpose, however a needle file can be also used with care.

Note that the input capacitor C1 uses the modified first inductor where it is soldered into place.

Remove sufficient of the inductor to accommodate the surface mount capacitor.

Note that the output capacitor C3 uses the modified circuit track from the Drain of GASFET F2.

Remove sufficient of the track to accommodate the surface mount capacitor.

Capacitors C1 and C3 are desoldered from the remaining LNB board, use the "white" capacitors used to couple the MMIC amplifier chips together or to decouple them.
For 5670MHz use (6cm) leave the stripline capacitor in place. For 9cm and 13cm remove enough of this capacitor to accommodate another "white" surface mount capacitor recovered from the remaining pcb (C2).

Keep the GASFET +Ve supply and bias circuits intact, but remove supply lines.

Drill 1mm holes close to the +Ve supply circuits (marked R and C in Figure 1) and counter-sink
them with a 4mm drill on the earth-plane side of the board to provide sufficient insulation for the +5 Volts supply lines that are fed to the supply circuits from under the board (P1 and P2).

Drill a 2mm hole as shown in Figure 1 for the -1.5 Volts bias supply.

Once this process has been completed, the box for the amplifier can be made up from double sided
copper clad board as shown in **Figure 2.** Solder the LNB two-stage amplifier to the long side of the box (solder both the earth-strip on the upper side of the board and the earth-plane lower side).

Suitable holes are drilled in the shorter sides as shown in Figure 2 to accommodate BNC sockets that are attached using 8BA (or equivalent) short length nuts and bolts. Note that the input and out BNC sockets are not symmetrical because of the LNB board layout. Add the BNC sockets and then solder the end walls together to form three sides of the box (both inside where possible and the outside as shown in Figure 2). Solder the BNC socket inner connectors to points X1 and X2 as shown in Figure 1.

Once the three sides of the box are soldered together, solder the earth-strip of the LNB to the "output" wall and solder the lower side's earth-plane to the these side walls where possible on the inside.

Remove the 7805 +5 Volts Regulator from the remaining LNB. Drill a hole in the longer side wall to accommodate the Regulator under the LNB two-stage pcb (i.e. the earth-plane side) and secure with a 4 BA nut and bolt. Drill suitably sized holes to solder feed-through connectors for the +Ve Supply (12v) and the –1.5 Volts bias supply (e.g.: a battery). Solder the input rail wire of the Regulator to the feed-through, the earth rail to the side of the box and the output rail to the +5 Volts supply rails run to the RC supply circuits to the GaAsFETs F1 and F2.

The Regular's input and output rails are decoupled to earth using 1000pF disc-ceramic capacitors. Solder a suitable length of wire to the feed-through connection point on the outside of the box.

Next add the GASFET bias rail. This is thin strip of pcb with the two 22k Ohms pre-set variable resistors soldered to it. The "variable" rail from each variable resistor being carefully soldered to the RC junction of the surface mounted bias circuit on the LNB. Solder the remaining variable resistor rail to the wall of the box. Solder the –1.5 Volts d.c. supply wire (insulated) to the pcb strip and pass this through the 2 mm access hole mentioned earlier. See Figures 1 and 2 for details.

Once the +/- Supply rails and the Regulator are in place, add the floor of the box that is also made from double-sided copper clad pcb.

Add the remaining wall of the box and solder wire tabs across the corners to ensure good connectivity.

Add a suitable wire to allow the –1.5 Volts to be connected.

Add two suitable length earth wire leads for the supply and bias.

Testing

Connect the amplifier's output (X2 BNC) to a suitable 13, 9 or 6cm receiver and connect the –1.5 Volts bias supply before the +Ve supply (12v was used).

The F1 and F2 Drain volts are set to about +3volts using a Digital Volt Meter to check the setting by adjusting the Bias variable resistors. (Check that F1 and F2 Drain volts can be varied from about +1.8 to +4 volts).

Connect a suitable aerial to the input (X1 BNC). Tune the receiver to a suitable signal source (e.g. beacon).

Adjust each 22k Ohms variable resistor for best maximum signal.

For use as a low power transmit amplifier, supply about 1mW of drive at 13, 9 or 6cm. Then adjust the 22k Ohms variable resistors for best output (one circuit gave about 80mW at 6cm.

In Use

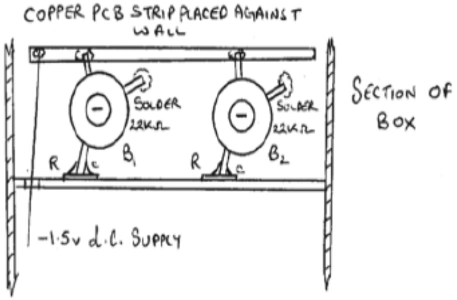
The design has been used from 13cm to 6cm on both the Rx and Tx systems for my transceivers.

Many QSOs have been made, including across the Channel, using these amplifiers as Rx and Tx amplifiers on these bands. For 6cm, two such modules are cascaded to give sufficient gain for the receiver.

A four-stage unit has been made up from two LNBs and tuned to act a multiplier from 1152MHz to 10368MHz, using two tuning pots made 13mm plumbing copper closed cylinders and copper 4 BA bolts to link stages.

FIGURE1: LNB BOARD CONVERTED FOR USE.

GASFET BIASING ARRANGEMENT



LNB MODIFICATIONS

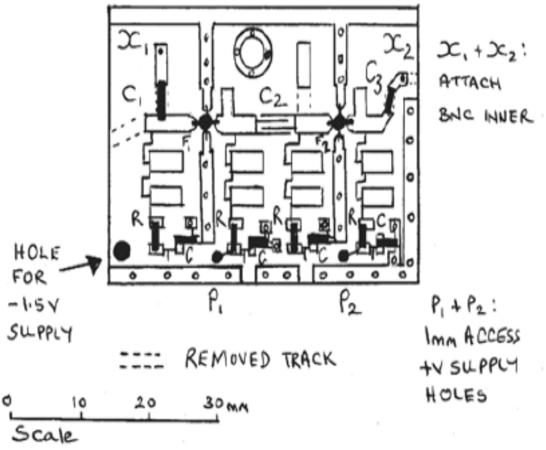
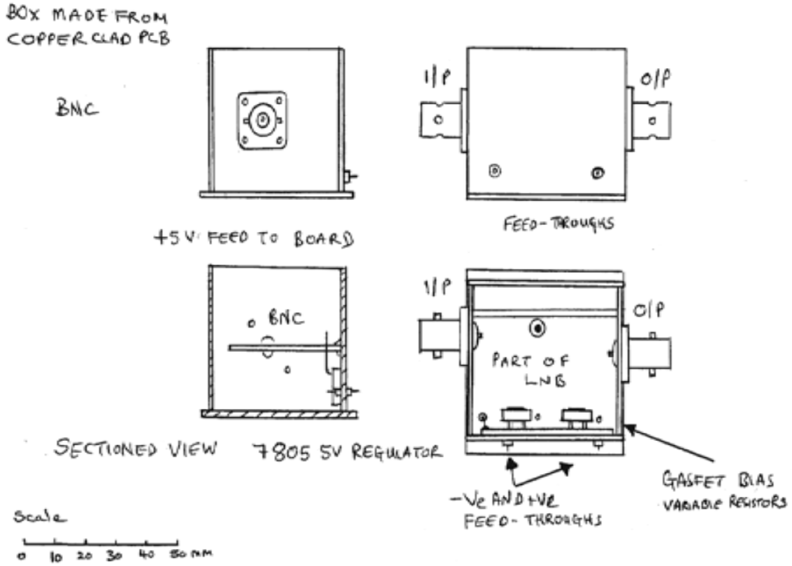


FIGURE 2: BOX CONSTRUCTION



23cm HEMT Tropo Preamp

Chris Bartram, GW4DGU

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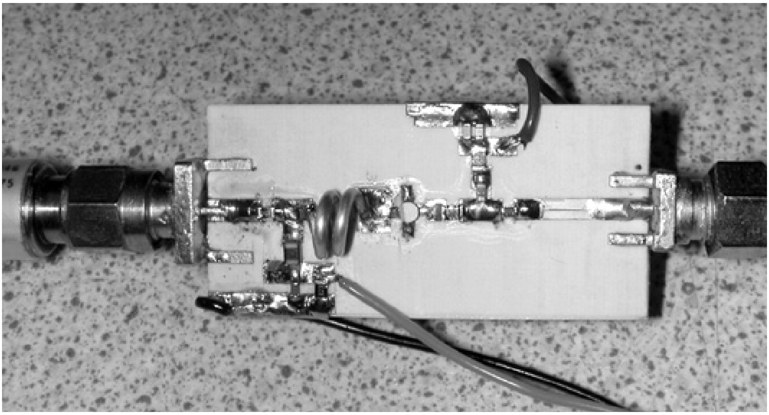


This preamp has been designed to be an excellent tropo preamp and a good second stage for EME. In practice, a pair of the amplifiers cascaded would probably be just about good enough for EME, but some redesign could probably shave 10K or more off of the noise temperature, at the expense of simplicity and possibly unconditional stability.

Please don't expect this to be a detailed constructional article. It's not. I'm trying to informally describe aspects of a project I've completed for my own entertainment and to possibly provide a small amount of inspiration for others. Experienced microwave equipment builders should find enough information here to duplicate the design but I hardly have enough time to earn a living, look after a small farm, and to play radio, let alone properly support a constructional project! If somebody wants to take the design further, and make PCBs available, I'd be happy to help, though.

I set-out to design a preamp covering the whole 1240 - 1315MHz range which would be easy to make with good, but not necessarily spectacular, noise figure, adequate gain, unconditional stability, good linearity and output return loss. The bandwidth of the preamp is very large. In gain terms, although not in terms of noise figure, it's still usable at 432! As it has good intermodulation performance, in many locations the amplifier could be attached directly to a 1296MHz antenna and not suffer from intermodulation and out of band issues.

Living in the primary service areas of two major TV broadcast transmitters, I can use the amplifier connected directly to a 1296MHz Yagi antenna without noticing intermodulation products but I may be lucky! In more stringent situations, a low-loss bandpass (or highpass) filter could be connected to the input. From a systems viewpoint, it makes good sense to separate the functions, rather than trying to design a narrowband preamp. If the filter and antenna have good return-loss at 1296MHz, the amplifier will still see something close to 50Ω, so the noise match will be unaffected and the noise figure of the filter-amplifier combination will be the sum of the preamplifier Noise Figure and filter loss. I'm working on the design of a simple-to-make, but very low loss bandpass filter, which I'll be putting ahead of this amplifier as a precaution. There are a number of options for a post-amplifier filter. My 'final' solution will probably be to use a second low loss input filter as an interstage filter, as my planned new transverter will have an excellent narrowband response from dielectric resonator filters.

Filters and antennas present impedances which can be far removed from 50Ω outside their passbands. For an amplifier to be stable with any combination of passive input and output load over the range of frequencies where the active device has gain is a highly desirable, but often difficult requirement to meet. It's even more difficult to prove in the real world. This preamplifier has been designed to be stable using the usual stability measures, and a detailed model been simulated to beyond 15GHz. It shows none of the usual signs of instability in simulation and I've so far failed to see any on the bench. That doesn't mean that there isn't a frequency somewhere in the spectrum where a combination of passive source and load impedances couldn't provoke instability. It just means that I've not yet found it!

Performance:

My prototype achieved the following performance at 1296MHz:

Noise figure: 0.55dB (T ≅ 40K)
Gain: 12.1dB
Input 3rd order intercept: -4dBm
Input 1dB gain compression: -11dBm
Output return loss: 20dB

The measured data agrees well with my simulations and also meets Bartram's First Law of LNA design: in the absence of linearisation circuitry, the input third order intercept of any low noise device, biased for low-noise operation, will be of the order of 0dBm. The plots were obtained from my VNA and written to a file via the IEE488 bus and then plotted using Open Office Calc running under Linux. The return loss graph looks a little noisy, as I accidentally read the data at 1dB resolution and didn't notice until I came to plot it.

My simulations also suggest that the noise figure remains good across the1.3GHz amateur band and is still OK at 1420MHz. I'd expect that to be the case, in practice, but I haven't yet measured the amplifier at other frequencies. Although apparently relatively simple, noise figure is a difficult parameter to measure accurately, particularly with modern test equipment! I am sceptical of many claimed noise figures. My measurement was made using a professional semiconductor noise source and my spectrum analyser, preceded by a low noise broadband amplifier. I guesstimate that the accuracy of my measurement is within about 0.3dB. A more accurate - and probably appropriate - method for amateur measurement of modern LNAs would be to return to the basic physics and use a low thermal mass 50 ohm termination dipped alternatively in melting ice and boiling water. However, that doesn't quite have the cachet of a £30k item of test equipment and it requires a modicum of understanding rather than the ability to read a display uncritically...!

Background

I have a number of Fujitsu FHX05 HEMTs in my component drawers following a successful EBay bid(!). OK, there are better devices, but not much better, and at £1 each.....?! Using Fujitsu's published device models, it was clear, following a lot of analysis, (using the Eagleware Genesys software

I use in my work) that it wouldn't be possible to guarantee unconditional stability out of band using a 'source feedback' topology. As an aside, although the source feedback topology has been around for twenty years or more, and is highly trendy - probably because it's possible to obtain a reasonable input match without degrading the noise figure - I've never found it possible to make a completely stable amplifier when using it. It's nearly always possible to find a region of instability, often at tens of GHz! I know of at least one other UK 1.3GHz preamp project (using Agilent PHEMTs with source feedback) which foundered for exactly that reason. For an LNA, good input match isn't actually necessary. Good input matching won't provide any more sensitivity (it's a matter of getting the right 'mismatch') but instabilities can completely wreck a potentially good NF.

The design that emerged after a number of simulations employs a conventional mismatched input circuit realised using lumped inductors for the input impedance transformation and gate bias feed and a shunt capacitance formed by a short length of microstripline. This forms a shunt-L, series-L, shunt-C network, which allows gate bias to be introduced at a relatively insensitive circuit node.

The HEMT source connections are grounded via 1mm diameter pins, cut and filed flush to the top (component side) surface of the PCB. This is critical for stability. The other component grounds are made by wrapping a piece of copper foil around the edge of the PCB at the appropriate places.

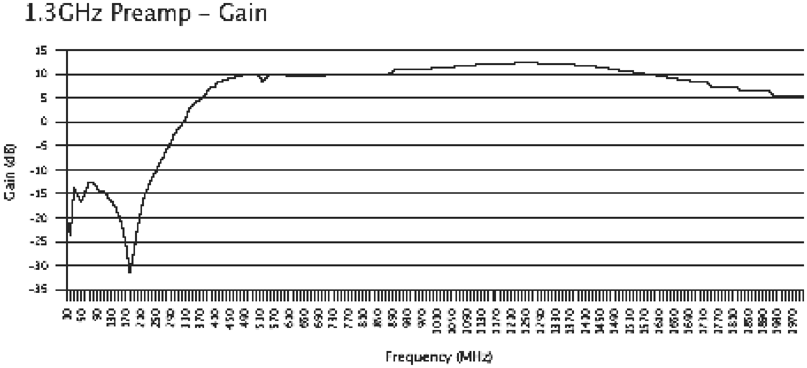
The output circuit is broadband. In order to control potential instabilities, caused effectively by the output resistance of the FET going negative at some frequencies, a small series resistor has to be inserted in the drain circuit. This causes a small degradation in the obtainable noise temperature but not as much as the device hooting away merrily at 30GHz! There is also a gain penalty, but gain is cheap, nowadays! The drain load is an inductor-resistor series network. This gives a good broadband output match.

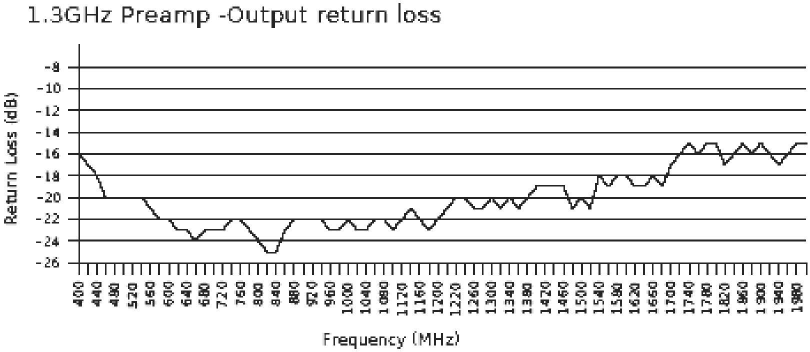
My prototype PCB was cut by hand from 0.75mm (0.030 inch) thick Rogers R4350 pcb material. I used just a scalpel, and a steel rule, working under a X10 binocular microscope. Using this technique I've made prototype LNAs to past 10GHz, and power amplifier boards to several GHz. R4350 is a low-cost, low-loss, PCB material intended for large quantity RF/microwave applications. It can apparently be processed by using standard production techniques employed for FR4 material. It's probably not part of the manufacturers design remit, but R4350 is also particularly easy to work by hand! I have generated a set of Gerber files for the PCB layout and put them on the <www.blaenffos.org> website. These may be used for non-commercial applications.

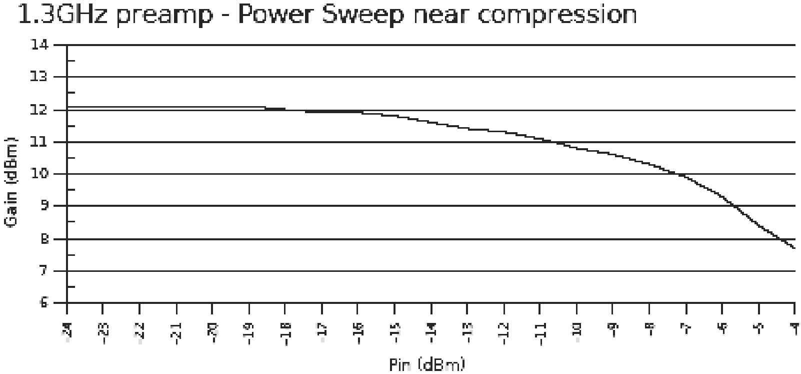
The drawing here has been through a number of format translations and is adequate as an illustration. It isn't to scale and a number of 'funnies' have crept in. I'd recommend that even if you intend to adopt the scalpel approach to PCB production, you download a Gerber viewer, such as PREVUE and use that to print scale drawings of the PCB.

Components

The most critical components are in the input circuit. The 6p8 capacitor really needs to be a low loss part. I used an 0603 AVX 'Accu-P' capacitor. Porcelain capacitors could be used, and even, at a pinch, a standard 0603 COG cap although that would have an effect on the obtainable NF. The 18nH shunt inductor is reasonably critical. A low-Q device could degrade the noise figure by some tenths of a dB. My


1.3GHz Preamp – Gain


1.3GHz Preamp -Output return loss


1.3GHz preamp - Power Sweep near compression

choice was a Coilcraft 0805CS series wound inductor. Perhaps the most critical part is the hand wound inductor. This is wound with a constant 2.5mm pitch on a 4mm mandrel e.g. a 4mm drill shank, such that the spacing of the centre of the start and finish of the winding is 5mm. There are no leads. Simply trim the coil so that it is a two turn helix. I used silver plated copper wire because I had some! Enamelled copper wire would also be suitable. In practice, unless high conductivity silver plating is used - and it's protected from corrosion, there's little advantage over copper. Don't even think of using tinned copper wire.

It may be necessary to slightly squeeze or stretch the inductor to optimise the noise figure, depending on exactly how the amplifier is built. If you feel this is necessary, thoroughly ground the gate of the HEMT before you do anything and remove the solder at one end of the coil. Make your modification, resolder and then remove the gate grounding. Otherwise you'll either lift a pcb track or damage the HEMT, or both! I know!

All other passive components were standard 0603 parts. The capacitors should have COG dielectric. In my case, the 5n6 inductor was a Coilcraft 0603CS wound part, but monolithic inductors would also be suitable. The FHX05 is the middle device in a series. It looks from the data sheet as though the devices are selected for noise figure in the 11GHz satellite television band.

The FHX04 has better guaranteed NF at that frequency, and the FHX 06, slightly worse. I suspect that there would be very little difference between them at 1.3GHz.

Enclosure

Although it might just be sensible to use a milled enclosure for a preamplifier built on a flexible substrate like ptfe/glass, that's really overkill. I tend to solder preamp pcbs into a lidless brass shim 'case'. This is more for physical protection than for RF screening, as I don't treat screening as a universal prophylactic! In this case, the bare PCB is small, the substrate material is relatively rigid and, providing the board isn't maltreated by being flexed, the amplifier will work entirely adequately just hung in the wiring! Flexing will break surface-mount components very easily and, even with a microscope, it's sometimes difficult to detect this visually.

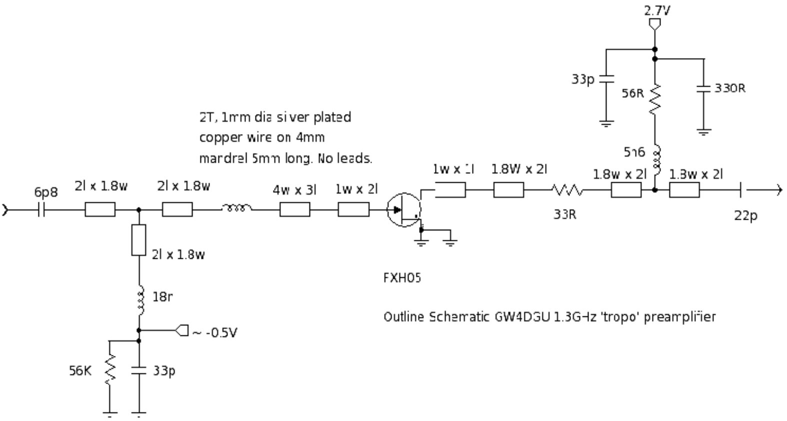
Although there are, of course, many situations where good screening is mandatory, it can bring its own problems. This is particularly true of amplifiers using devices with bandwidths of tens of GHz.

There are hazards in packaging microwave amplifiers: putting-on lids leads to many potential problems as it is frighteningly easy to excite cavity resonances. Identifying and killing these can be as much work as designing the amplifier in the first place! HEMT precautions It's very easy to damage HEMTs and not realise it. They are extremely susceptible to static damage during assembly and from supply line transients in operation. This doesn't usually show as a significant change in the dc parameters, just as a change (for the worse!) in noise figure. Be VERY careful and take extreme precautions to avoid static damage when soldering the device into circuit.

Power supplies

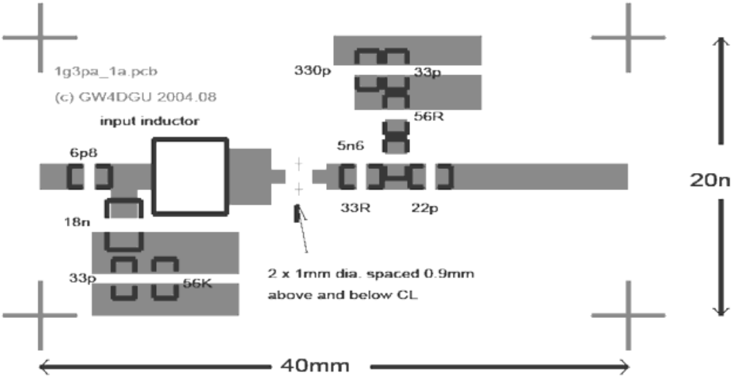
The preamp requires a drain supply of +2.8V, 10mA. The gate bias required for this will be of the order of -0.6V. As HEMTs, along with most LNA devices are susceptible to supply-line transients, I now operate my preamps from isolated power supplies. In this I use a cheap packaged transformer-coupled inverter which converts 8 - 36V dc into ±5V, this largely isolates the preamp from supply line transients, and lets me power the preamp from my 28V antenna relay supply, as I energise the antenna c/o relays on receive. Following the inverter, I filter and clamp the ±5V lines and use these to power active bias networks. I'll write-up this 'preamp power supply for the paranoid' in the near future, along with details of the sequencing circuitry I use.

Editor's note: This article (and the preamp design) is copyright material. Permission to reproduce it in any form must be obtained from Chris Bartram, GW4DGU.
He is accessible by email at: gw4dgu@blaenffos.org



Outline Schematic GW4DGU 1.3GHz 'tropo' preamplifier

PCB

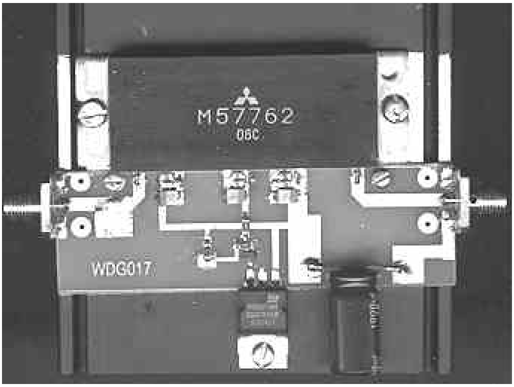


Chapter 5

Power Amplifiers

TheWDG017 20W 1.3GHz Linear Amplifier

by Charles Suckling, G3WDG

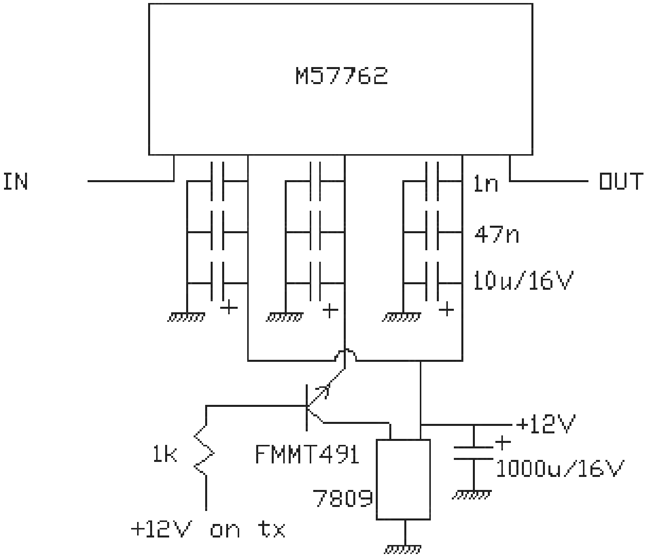


Editor's note
This design is now obsolete as the module is no longer available, but these amplifiers and modules appear from time to time on the surplus market , so worth publishing as an example of a good design before we move on to the module's replacement.
The WDG017 20W PA uses the well-proven Mitsubishi M57762 power module which is designed to cover the whole 1.3GHz band. The power seems to peak around 1260MHz and the module still works very well at 1296MHz. The WDG017 employs a transistor switch to enable the PA to be shut down during receive periods, to save energy and lower the heatsink temperature.

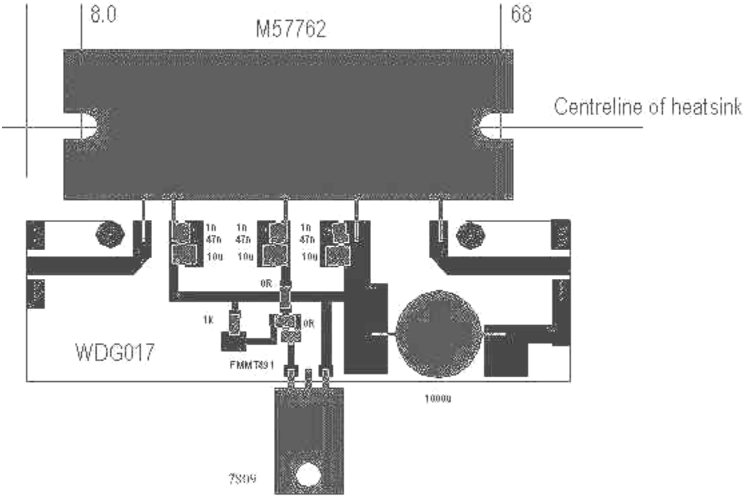
Construction
The original prototype was constructed using a commercial heatsink as the chassis. The heatsink measured 80 x 100 x 30mm (Farnell 738-906) and the construction notes refer to the use of this heatsink. Other prototypes have been built on a 6mm (or 0.25") thick aluminium plate as the chassis.
This has some advantages, in that any heatsink out of the junk-box can be used (provided it is big enough!), or the plate may be water-cooled. In these cases the plate chassis is just screwed on to the main heatsink. Also, most aluminium plate is easier to work than the metal used to make heatsinks, which tends to be rather soft, and there is no need to scrape off the anodising (a rather tedious operation!).

1. For 1W PA, cut heatsink to size (45 mm). 18W PA uses full size heatsink.
2. Drill and tap holes for securing the module to the heatsink. Refer to assembly diagrams for location of holes. Drill 2mm and tap M2.5 (1W), drill 2.5mm tap M3 (18W).
3. Temporarily, fit module to heatsink.
4. Locate pcb onto heatsink - refit and tighten module screws so that pcb aligns to edges of heat sink without

circuit diagram



layout



under/overlap and module pins align to tracks on pcb.

5. Temporarily, sellotape pcb in place and drill through holes in pcb either side of RF lines part way in to pcb (to make drill location holes).

6. Drill through heatsink in marked places 1.6mm diameter and tap M2. Drill carefully, with lubricant, to avoid breaking drill. Also tap with lubricant.

7. Open holes previously drilled through in pcb to 2mm dia.

8. Fit pcb temporarily to heatsink with M2 screws.

9. Locate voltage regulator and drill 2.0mm hole and tap M2.5 for its fixing screw

10. Locate connectors and drill/tap heatsink for retaining screws. Use 2 x M2 if using SMA types.

11. Skip 12-14 if using plated through hole pcb

12. File ends of heatsink to allow clearance for ground pins

13. Drill through grounding holes in rectangular pads near module leads to mark heatsink (6 places) and drill these out a few mm into the heatsink to sufficient diameter (4mm suggested) to give clearance to ground pins and associated solder fillet).

14. Fit veropins to pcb, heads on ground plane side, cut to leave about 1mm protruding on pcb side,
and solder to both sides. File excess solder off heads on ground plane side. Keep solder fillet within confines of blind relief holes or scrape off excess so board still sits flat on heatsink.

15. Scrape anodising off heatsink under pcb and module area.

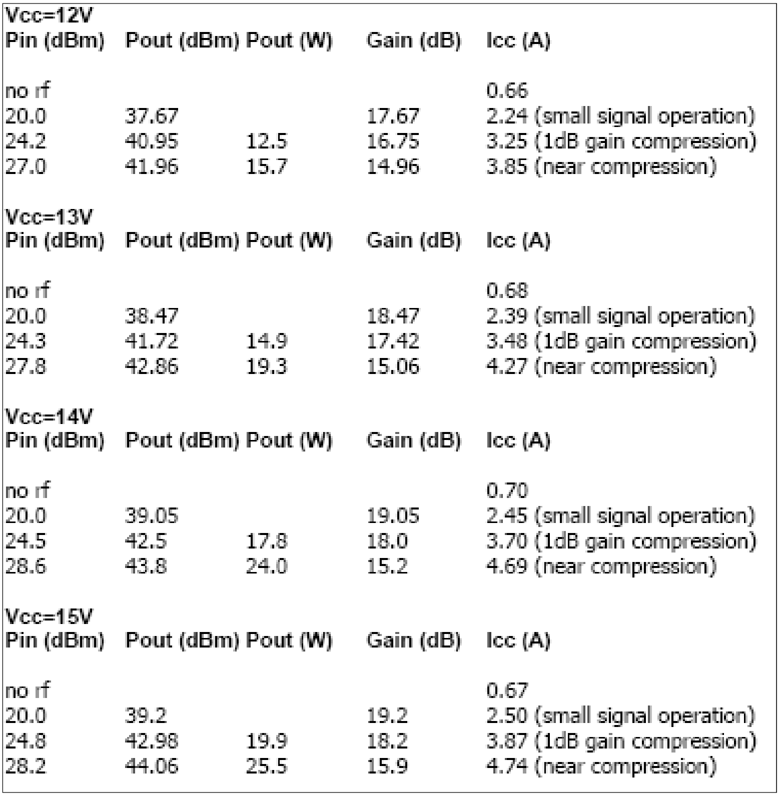
16. Reassemble pcb to heatsink and fit connectors. Solder connectors to ground strips on track side of board using decent sized iron. If you have only a small iron, preheat the heatsink to 150C approx and while hot make the solder joints.

17. Mount other components including the module to pcb.

Testing and using the PA

Apply volts, apply drive and you should have output! Performance of one of the prototypes at 1300MHz is shown below. The data sheet for the M57762 quotes an absolute maximum supply voltage of 17V, and that the device should survive a 16:1 VSWR with a 15.2V supply at 18W output power. For most applications, a supply voltage up to 14V is recommended. If equipment is available to measure the input VSWR, a small improvement may be possible by adding a tuning tab to the input line (see photo at the beginning of this article).

It is recommended to shut down the PAs on receive. This can be done by removing the 1k resistor and applying +12V on transmit via a 470R resistor to the pad previously connected to the bottom end of the 470R resistor. On receive, the lack of a voltage effectively shuts off the module.



Vcc=12V

Pin (dBm)	Pout (dBm)	Pout (W)	Gain (dB)	Icc (A)
no rf				0.66
20.0	37.67		17.67	2.24 (small signal operation)
24.2	40.95	12.5	16.75	3.25 (1dB gain compression)
27.0	41.96	15.7	14.96	3.85 (near compression)

Vcc=13V

Pin (dBm)	Pout (dBm)	Pout (W)	Gain (dB)	Icc (A)
no rf				0.68
20.0	38.47		18.47	2.39 (small signal operation)
24.3	41.72	14.9	17.42	3.48 (1dB gain compression)
27.8	42.86	19.3	15.06	4.27 (near compression)

Vcc=14V

Pin (dBm)	Pout (dBm)	Pout (W)	Gain (dB)	Icc (A)
no rf				0.70
20.0	39.05		19.05	2.45 (small signal operation)
24.5	42.5	17.8	18.0	3.70 (1dB gain compression)
28.6	43.8	24.0	15.2	4.69 (near compression)

Vcc=15V

Pin (dBm)	Pout (dBm)	Pout (W)	Gain (dB)	Icc (A)
no rf				0.67
20.0	39.2		19.2	2.50 (small signal operation)
24.8	42.98	19.9	18.2	3.87 (1dB gain compression)
28.2	44.06	25.5	15.9	4.74 (near compression)

Precautions using Mitsubishi RF Power Modules

Charlie Suckling, G3WDG

Many stations are using RF power modules very successfully, but on occasions there have been problems where modules have failed or degraded. The most likely cause for this is overheating, particularly where modules are driven hard or are operated continuously.

The extract reproduced below may be useful regarding this and other general precautions. It is important to ensure that the heatsink surface is flat and free from burrs, and that the heatsink is adequate to dissipate the heat generated by the module. Also note that Mitsubishi recommends the use of thermal grease to improve the heat conduction from the module. Mitsubishi states that the temperature of the module flange should not exceed 90C or irreversible damage may occur.

I have recently had occasion to dismantle a degraded M57762 and found a cracked chip capacitor in the output network and some fried thick-film resistors, so this may well be true! Under high drive conditions, the M57762 module, used by many stations on 1.3GHz, may dissipate over 40W of heat. The heatsink specified in the article in Microwave Newsletter above may not be large enough for continuous use. Also note that for optimum operation, the heatsink should be mounted with its fins vertical so that air can flow freely and achieve optimum cooling. Measurements show that mounting the heatsink flat with the fins downwards will increase the thermal resistance by as much as 50%!

It has been found that as supplied the modules are slightly bowed, which means that the centre of the heatsink may run considerably hotter than it does at the ends, near the mounting screws.

Measurements with a thermocouple have shown that centre of the module flange seems to run some 20-30 degrees hotter than the heatsink to which the module is bolted, suggesting that the heatsink itself should not be allowed to run hotter than about 60C. The temperature of the flange near the mounting screw only runs 10 degrees or so hotter than the heatsink and may lead to an optimistic view of the temperature of the flange where it really matters!

Mike Willis G0MJW has come up with an interesting suggestion to prevent operation continuing if the heatsink gets too hot. This is to use a bi-metallic thermal switch bolted to the heatsink to shut down the module if the heatsink gets too hot. Once the heatsink cools, the switch resets automatically and the PA can be used again (and the overs kept shorter!). Suitable switches are available from Farnell and RS at relatively low cost. One option, suggested by Andy Talbot, G4JNT, is to used forced air cooling of the heatsink. This should enable a physically smaller (and cheaper) heatsink to be used.

Various small fans are available quite cheaply these days (e.g. computer cooling fans). This is an excellent suggestion for situations where air can move freely around the PA.

In applications where two modules are mounted on a common heatsink for combining, excessive temperatures are even more likely and great care needs to be taken in selecting an adequate heatsink.

Mitsubishi 23cm ‘Brick’ PA Replacement

Gleanings from the Internet Reflectors

When Mitsubishi introduced the replacement 23cm power module, some interesting discussion took place on the WA1MBA microwave internet reflector. The emails refer to the ***RA18H1213G module***

From Sam G4DDK

Mitsubishi 23cm PA module

I mentioned in an e-mail last week that I had ordered one of these new Mitsubishi 23cm PA modules from the UK distributor, now that they are in stock. I have received my module and built it into a suitable heatsink case and have some preliminary results.

I am currently achieving 25W saturated output from 13.5V supply and Vgg set at 4.6v. This is about 0.1V above the apparent recommended Vgg value. I will experiment with the bias level when I can get the amplifier onto IMD testing. 4.6V bias gives 1.7A quiescent current. At 25W output the supply current is about 8A. The required drive is about 500mW for this output power.

A few observations are in order. The power leads are not thick enough (or maybe too long - they need to reach my only available PSU!) and so I get about 0.7V drop at the amplifier at 8A. When I shorten these I expect to get about 1dB more output. At this level the efficiency is about 25 percent.

When switched to transmit but no RF drive the quiescent current slowly increases to about 1.8A and then stabilises. After several seconds of hard RF drive and then removing the drive the result is the quiescent current slowly falling back to about 1.7A over several minutes. The heatsink is more than adequate to dissipate several hundred watts for a reasonable temp rise above ambient.

I am surprised at how much drive the amplifier needs. I haven't analysed this yet, but initial reaction is that it is more than expected even allowing for the saturated output condition. I will have to re-think the transverter drive chain levels!

I have posted a few pictures on my web page at **www.g4ddk.com.** Follow the links from Technical information.

One well-known amateur radio equipment producer has posted information on his commercial single and dual module amplifiers using the Mitsubishi module. The claims appear to be that two bricks will give 90W saturated output. Maybe there is an even more powerful version of the module available now?

I am pleased with the initial results. I haven't blown up the brick yet despite a pessimistic comment regarding the new LDMOS amplifier modules!! 45W? Ummmmm!

I have had a number of responses to my comments about the new Mitsubishi module. I'm responding to the group instead of to individuals. I think the information may be of general interest. Now that I have further examined the 'full' data sheet mentioned by Jeffrey Pawlan a few days ago I think it is now likely that I have been very cautious with the module (nothing wrong there!) and that the quiescent current can be increased beyond the 1.7 amps I'm now using. The data sheet suggests it should be possible to get 40W sat. out at Vgg of 5v, corresponding to something like 4A quiescent! Increasing the VCC to 15V may be how to achieve 45W but I can't help thinking that could be expensive. The data sheet shows 200mW drive should be sufficient to achieve 30W output and that 300mW is the maximum so I will definitely need to back off from the 500mW tried. My tests were all done at 1296MHz but a brief check showed a little more gain available mid band and useable gain to 1325MHz (the top of the band in the UK).

Does anyone have any new information on the LDMOS bias 'memory effect' mentioned in one or two professional LDMOS HPA papers? A colleague, who was working on these devices for cellular amplifiers until a year or so ago, thought that the concerns were overstated.

I will continue my tests later this week. I am extremely hopeful that these modules will help to 'drive' interest in the band. LDMOS does look like the way forward for those of us who don't want to use valve amplifiers for various reasons and yet want to run more than the usual 10W bricks.

BTW, it does seem that these modules may be somewhat cheaper to buy in the southern hemisphere leading to thoughts of cost-effective multi-module amplifiers without it costing an arm and a leg!

From Grant G8UBN

The data sheet says that the absolute maximum O/P power is 30W so I wouldn't recommend trying to get more than 25W out - they are designed to be 18W devices. Also, the devices are well into saturation at this point; fine for CW (or FM ATV if you have a big heatsink) but not very desirable for SSB.

I've built an amp with this module but haven't had time to test it yet, but the quiescent current was 3.5A@13.8V. I don't recommend going above 13.8V. I'll do the RF testing when I get time - too busy with other things at the moment. LDMOS devices definitely have a 'memory' effect, although often it manifests as a 'drift'. Some manufacturers of mobile phone (cellular) base stations go to extraordinary lengths to ensure that the long term (several years) bias conditions are maintained; this is essential to maintain linearity for the new generation systems such as EDGE and W-CDMA (aka '3G').

However, I am not sure if the Mitsubishi modules are LDMOS - they are certainly enhancement-mode MOSFETs, but I'm not convinced that they are laterally diffused (LD). If they are, then this would explain the poor efficiency - most, if not all, LDMOS FETs that work above 1GHz will only work properly at ~26 -28V DC, and I believe that there is some development in 48V LDMOS devices. Running them from 12V gives poor performance - which would explain the high quiescent current of the RA18 .... if it is indeed LDMOS.

As for combining them - it should work OK as long as the splitters/combiners have good isolation between ports; Wilkinson, Gysel, 90 degree hybrids and 'Rat-races' can all exhibit good enough isolation if designed and built properly.

I'm working on a design for a combiner/splitter that can be configured as a Wilkinson for the input and a Gysel for the output; initial results look promising but there's a bit more tweaking to do.

More information on the Mitsubish Power Amplifier Modules

Sam Jewell, G4DDK

These Measurements are the results I obtained in my own shack workshop. Please note that these results are for my module, operating at 13.4V and with the bias set to 5V. The module heat spreader has not been sanded flat and the module is mounted to its (substantial) heatsink with two screws into the heatsink and a small amount of standard silicone heat sink compound used between the module and the heatsink. Prolonged use in this way has not resulted in any failure (so far!). No screen has been used over the top of the module and the output connections are via UT141 coaxial cable as shown on my web page at: **www.g4ddk.com**

Test System

The test system consisted of one R&S SMG signal generator and one HP 8640B signal generator combined with a Hatfield Instruments 3257-03 hybrid combiner.

The combined generator output was carefully measured on the R&S FSB spectrum analyser to ensure 3rd order products were <-46dBc at 144MHz.

The separation of the two tones was 100kHz (1296.250 and 1296.350MHz).

The resulting 40mW PEP 144MHz IF signal was fed into my new homebrew 23cm transverter.

A Mini Circuits ZHL42 broadband amplifier was connected at the output of the transverter and with the transverter TX IF input attenuator carefully adjusted to give 40mW PEP at the ZHL 42 output the 3rd order products were measured at -42dBc. The transverter couldn't quite achieve this good an IMD performance at it's rated 50mW output, hence the assistance from the ZHL42 in order to reduce any contribution from the transverter IMD.

The ZHL42 40mW PEP output was then connected to the Mitsubishi amplifier and the RA18H1213G amplifier output power measured on an HP435 power meter with HP8481B sensor and Weinschel 30dB/50w attenuator. The indicated power was 10W. Since the amplifier was driven by two tones at equal level, the output PEP was 20W.

Results

When measured on the FSB the 20W PEP output spectrum showed:

- the 3rd order products to be -17dBc,
- the 5th order about -24dBc.

At 16W PEP output

- the 3rd order products were -20dBc
- the 5th order -30dBc
- the 7th order -42dBc.

Higher order products were well suppressed.

Conclusions

My conclusion is that the amplifier, whilst not outstanding on 3rd order, has very acceptable 5th and higher order IMD performance and should sound very clean on air (assuming it is used with a sufficiently clean IF rig and transverter......).

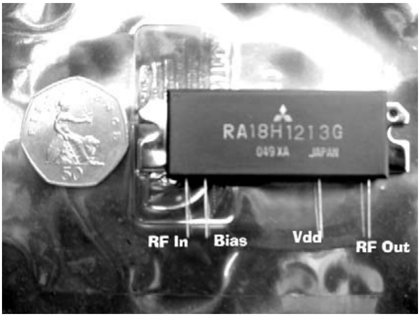
Note that this is a single amplifier measurement. Varying the operating conditions could and probably would cause significant changes in the IMD performance.

I am happy with what I've seen and I hope this will go some way towards redressing the balance about these new amplifiers. Some previous posts have been quite pessimistic.

As always, your mileage may vary. I hope this information will be of some interest to other 23cm band users.

Using Mitsubishi Power Amplifier Modules

Grant Hodgson, G8UBN



Introduction

There have been various discussions on the US microwave reflector regarding the use of the new Mitsubishi RA... series power amplifier modules as amplifiers for both 1.3GHz and for VHF/UHF talkback. This article is intended to explain how to get the most out of these very useful and versatile devices and to counter some of the myths and half-truths that have developed over the years about PA modules in general.

This article may seem rather long but it is surprising how such an easy-to-use component as a 12V 50 ohm PA module can be abused and even destroyed though lack of understanding of how these devices should be handled and operated. So what follows is an attempt to explain in detail some of the more subtle points.

History

The RA… series of Power Amplifier modules from Mitsubishi follow on from the hugely successful M5/6xxx series which were introduced in the 1980s as a means of easily amplifying RF power in the VHF and UHF frequency ranges. Literally millions of M5/6xxx series PA modules were sold worldwide to both commercial and amateur radio customers. They offered many advantages compare to RF amplifiers using discrete power transistors and everything was included in a single module – all that was required externally was a heatsink, a main DC supply (which could be taken directly from a 12V car battery), a regulated bias supply and some de-coupling capacitors.

Output powers ranged up to 60W at 144 and 432MHz and many of the modules could be used for SSB, although not at the full rated output power.

The M5/6xxx series are now obsolete; the last ones rolled off the production lines around 2003/2004. Many of these PA modules are still available, but as time goes on it is getting ever more difficult to find them, as the old stock gets used up.

New RA… series PA modules

One of the main problems for Mitsubishi with the M5/6… series was that the modules did not have a very wide frequency range and consequently, in order to meet the needs of thousands of different customers, together working at nearly all frequencies from 50MHz to 1.3GHz, many different modules had to be produced. In fact, at one time there were more than 200 different types of Mitsubishi PA module in production, each one with a separate part number and the overheads involved in managing such a huge number of relatively similar devices were enormous. Mitsubishi's solution to this problem was to scrap the old M… series entirely and develop a range of new modules which would offer a much greater frequency range, thus reducing the number of variants. One of the knock-on effects of this was that, generally speaking, the new RA.. series modules are cheaper than their M5/6xxx counterparts.

The technology used for these new modules still uses silicon (as opposed to other semiconductor materials such as Gallium Arsenide amongst others) but, instead of using bipolar transistors, the new modules use MOSFETs. MOSFETs offer a number of significant advantages over bipolar transistors, including wider frequency range, higher gain, higher power output, better linearity and practically no bias current. The supply voltage remains the same, which means that some applications can use the new modules with only a small number of changes, which will be described later. Note that the new modules do not use a technology called 'LDMOS' (Laterally Diffused MOS) – this is a slightly different type of technology, which usually requires a 28V DC supply. The MOSFETs used in the RA... series modules are 'enhancement mode', which means that a +ve gate voltage is required for them to operate, and zero gate voltage completely turns them off.

The modules that will be of most interest to radio amateurs are the high power modules, and these are the ones that will be discussed in more detail.

Part numbering

The M5/6… series of PA modules simply took numbers from a list, in sequence, as they were developed. It was impossible to determine the performance of a module simply from the part number; one had to have access to the data sheet or have a very good memory (think of how many devices there were!) .

The new RA series have part numbers that relate directly to the device's performance. Therefore, it is possible to ascertain many of the operational parameters without access to data sheets, and conversely it is possible to find the part number of a module from a specification.

The new part numbers are of the form :

RAaabccdde

RA RF PA Module
aa nominal ouput power in Watts
b supply voltage code
cc lower frequency limit
dd upper frequency limit
e Frequency Multiplier – either M or G.

The supply voltage code is as follows :-

H 12.5V
N 9.6V
M 7.2V

If the frequency multiplier is M, then the lower and upper frequencies are in 10s of MHz. This gives the operating frequency to within 10MHz, the exact operating frequency limits are specified on the datasheet.

There is only one module with a 'G' suffix, with is the RA18H123G.

Examples: RA45H4047M
45W, 400 – 470MHz, 12.5V
RA30H0608M
30W, 68 – 88MHz, 12.5V
RA03M8894M
3W, 889 – 940MHz, 7.2V
RA18H1213G 18W,
1240 – 1300MHz, 12.5V

Note that in practice the modules can be used outside of their stated frequency range, for example the RA30H0608M can be used at 50MHz, with good performance.

Using the new RA series modules

None of the RA series modules are a direct replacement for any of the M5/6… series modules.

There are a number of detail changes that the user must be aware of, in order to avoid the possibility of damage to these modules.

Static Discharge

All electronic components that use FETs are susceptible to damage from static discharges, and as such the same precautions must be taken when handling the new PA modules as when handling GaAsFETs and similar devices. The older M5/6xxx series used bipolar transistors and as such were not easily damaged by static. In addition to the possibility of damage from static, the modules can also be damaged by small leakage currents from the tip of a soldering iron. Therefore, it is imperative that any soldering iron used has a ground connection to the tip, and that this connection is checked with a multimeter – the author has managed to destroy at least one module due to a fault with the earth connection on a high-quality (Metcal) soldering iron; a fault that took a lot of time to investigate but was easily corrected.

Package and Pinout

All of the new high-power PA modules operate from a nominal +12.5V supply, and can be operated at 13.8V without damage. The package and location of the pins are the same as previously, which is very convenient. However, the new modules have only 4 pins instead of 5 for the M5/6xxx series.
The pin connections are :-

Pin 1 RF in
Pin 2 Bias supply
Pin 3 Main DC supply
Pin 4 RF out

Note that the bias connection has moved from pin 3 to pin 2. Therefore, if a PCB is being used which was designed for the M5/6xxx series modules, then the PCB will require modification, although this will usually be a simple track cut and wire link.

Bias supply

This is probably the biggest difference between the M5/6xxx series and the RA… series. Most of the M5/6xxx modules required a bias supply of 9V, although there were some exceptions where an 8V supply was needed. The bias current for the M5/6xxx series varied from module to module but as an example the M57762 used on 23cms required approximately 700mA, which would typically have been supplied from an L78S09CV or similar voltage regulator in a TO-220 package.
The RA… series modules require a bias voltage of between 4.2 and 5V, and the bias current is tiny - in the order of 1mA or so. Therefore it would be possible to use much smaller, lower-current voltage regulators, although high current regulators will still work of course.

With no bias supply, the FETs are completely turned off, and the PA module acts as an attenuator. Therefore, it is possible (and desirable) to connect the main DC supply to a constant +12V supply; there is no need to use a separate high current relay or switch. PTT operation can then be accomplished by keying the bias supply.

The setting of the bias voltage is the subject of some discussion. Unlike the M5/6xx series PA modules, there is no one, single optimum bias supply voltage. In fact it is possible to set the bias voltage to suit individual requirements. The Mitsubishi data sheet is very vague in this area, although the device parameters have been characterised with a bias voltage of 5V. However, many users report good results with bias voltages lower than this, and in one case as low as 4.2V. What is known is that the drain current, output power and gain increase as the gate voltage increases,particularly above 4V. Also, the RF/DC efficiency of the modules decreases as the bias voltage is increased, due to the fact that the drain current increases more rapidly than the output power.

Decoupling and PCB

The supply pins require decoupling in order to reduce the possibility of instability. Mitsubishi recommend a 4.7nF capacitor in parallel with ‘at least 22uF’. The 4n7 needs to be placed as close to the PA module as practicable. This is easily accomplished, especially since there are only 2 DC supplies.

The gain of the RA… series modules is much greater than the M5/6xxx series. The modules have been designed to be stable into any output load with a VSWR of up to 3:1. However, this assumes that the supply pins are sufficiently de-coupled. In order to achieve the best decoupling performance, a very low impedance path must be provided to ground. The best means of achieving this is with a printed circuit board that uses plated-through holes. These PCBs are more expensive to produce than 'non-PTH' boards, but give much better performance than boards where the grounding has been achieved by other means, especially at 1.3GHz. This is one case where it really is worth spending a little extra in order to achieve the best performance. One further note which is related to both de-coupling and static discharge – it is worth placing a high-value resistor in parallel with the 2 DC supply pins and the RF output pin in order to further reduce the possibility of damage due to static discharge. This is best done by soldering the resistors to the PCB before the module is fitted; that way the pins of the module will be grounded as they are soldered. The author uses 68k, but any value from 10s of kohms to 1Mohm could probably be used. Note that the RF input pin is internally connected to ground with a PI attenuator, and does not require an external resistor.

Grounding

There is some considerable confusion as to how the PA modules are grounded. The modules are designed to be mounted on a large, flat heatsink. With theM5/6xxx series modules, the best thermal performance is achieved by applying a thin, even layer of thermal compound on the flange of the heatsink. The application of the thermal compound means that a good electrical contact between the flange of the module and the heatsink cannot be guaranteed. Therefore, the electrical ground path for both DC and RF from the module to the PCB is via the mounting screws and the heatsink! This may seem totally counter-intuitive, but this is indeed the situation, which leads to some interesting points that need to be considered :

- The mounting screws should be made of brass and be bright zinc plated; screws made of other materials such as steel will not give as good performance and should be avoided.
- The mounting holes in the heatsink must be tapped; it is not sufficient to drill clearance holes in the heatsink and use long screws with a nut on the other side.
- The mounting holes must be clean – this means that any residual cutting fluid must be removed with a cotton bud or similar, and great care must be taken to ensure that no thermal compound gets on the screw thread.

(The author is aware of several 1.3GHz amplifiers that were unstable and actually oscillated; in all cases the instability was cured simply by cleaning the mounting screws.)

Black anodised heatsinks can be used without the need to remove the anodisation from under the module.

Chapter 6

Accessories

Easy 3db splitter for 10GHz

by Tim Leighfield, G3KEU

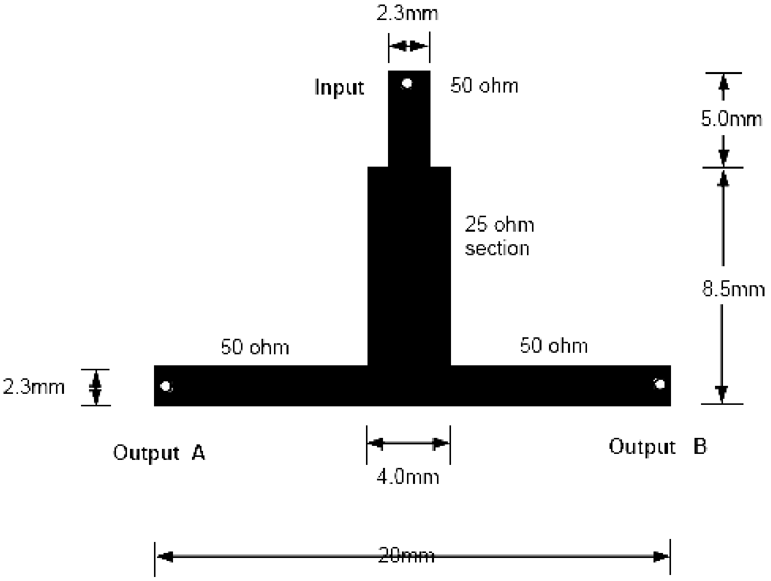
Occasionally the need arises to divide the output of an RF source to feed, for example, the 10224MHz LO inputs of separate receive and transmit mixers. Less common is the need to split the received signal for distribution to two receivers.

In both cases, a 3dB loss is obviously incurred when comparing output to input levels.

The PCB layout shown here (not to scale) can be etched on RT 5870 Duroid. If other types of microwave printed circuit board are used, the dimensions shown below would need to recalculated.

With care, the lines could be cut, rather than etched, using a sharp scalpel and a hot soldering iron to encourage the surplus copper laminate to lift. Once etched or cut, the pcb should be soldered into a small tin-plate or brass box.

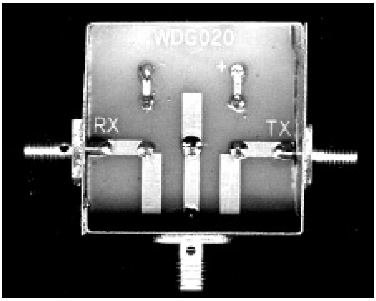
SMA connectors can then be soldered to the box and directly to the ends of each stripline forming the arms of the divider.



23cm Accessories

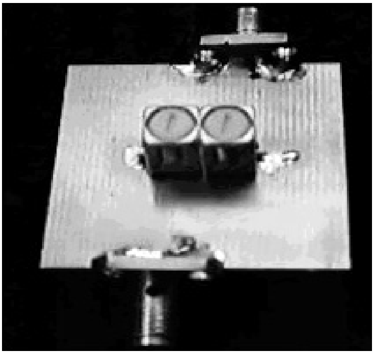
Useful ‘add ons’ from G3WDG

TX/RX RF switch using a "low cost" pcb mounting relay:
A tx/rx relay for 1.3GHz has been prototyped using a relatively low cost pcb relay. The performance of the relay has been enhanced by tuning out the reflected power with tuning tabs on the FR4 pcb. Insertion loss of the relay is about 0.3dB, isolation is better than 40dB and return loss in “through mode” is around 20dB. The relay is rated to 15W power handling (To be tested).



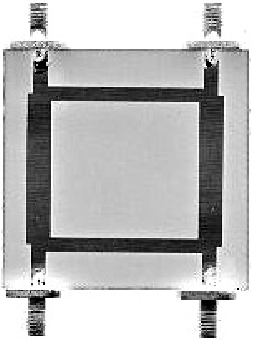
An add-on filter using a Toko helical filter for the 23cm transverter:
A simple filter using a TOKO helical filter has been developed as an "add on" to the 23cm transverter to clean up the spectrum if required by local regulations.

Insertion loss is only 1.8dB, and the filter provides about 32dB and 42dB local oscillator and image rejection respectively.



Quadrature Couplers:
Various quadrature couplers are being worked on, as splitters and combiners for balanced PAs. These use the branched arm configuration which theoretically produce two signals. -3.0dB down on the input signal, 90 deg apart in phase.

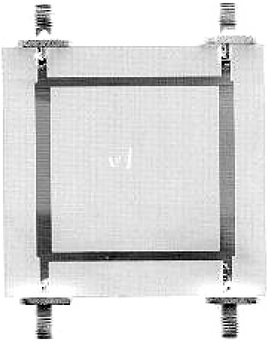
Non ideal couplers do not have an equal amplitude split and may not achieve 90 deg phase difference. For balanced amplifiers, an amplitude difference of 0.1 to 0.2dB is usually tolerated, and phase inaccuracy of up to about 5 degrees from true quadrature is generally acceptable.
To date, two microstrip couplers have been developed, shown in the photo above.
The one on the left uses 1.6mm thick FR4 as the substrate material, while the one on the right uses 0.8mm thick duroid.



Performance
FR4 coupler:
S11 -35dB at 1296
S21 -3.45dB at 1296 phase = -140
S31 -3.50dB at 1296 phase = 129
S41 -35dB at 1296
S21-S31 0.05dB phase = 91 deg, dissipative loss = 0.47dB. Estimated power handling 40-50W

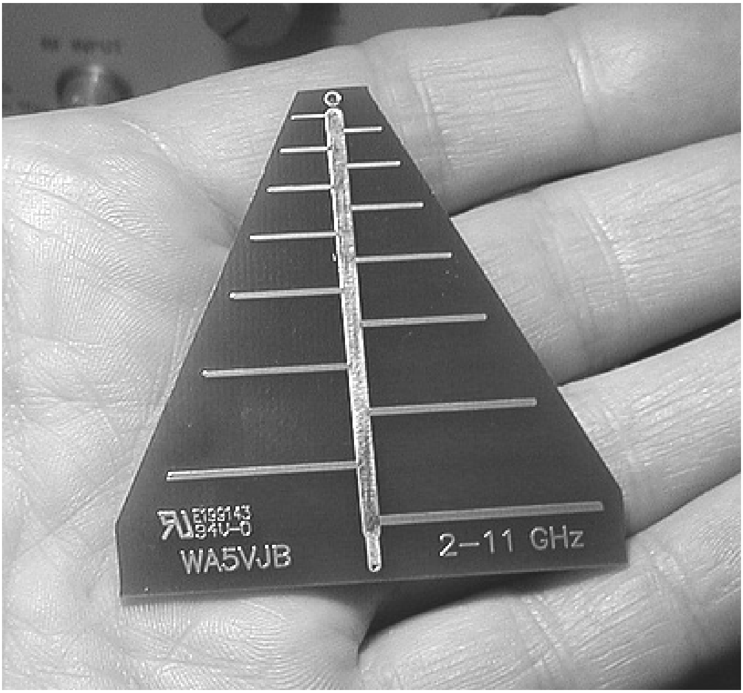
Duroid Coupler
This coupler covers the whole 1.25 to 1.35GHz band with excellent phase and amplitude balance. It should be an excellent candidate to combine the output of two 35W amplifiers.

Loss is about 0.15dB, and power handling should be in the region of 80W. Input return loss (IRL) is better than 20dB (1.2 VSWR over the whole band and reached about 38dB at 1296. Isolation at 1296 is about 40dB.



An RF Microwave Signal Sniffer

by John A. Jaminet, W3HMS



Often, I like to know immediately if my transverter or amplifier is QRV… 100% …YES or NO !

In the past, I have found that it‘s possible to see a signal on a voltmeter from the MON jack on my DB6NT transceivers …. without in fact having an emitted signal. The fault was my antenna relay! Thus I made a promise to myself that I would eliminate this circumstance once and for all. The means which I used is simple, effective and quite old in the history of radio.

It is a field strength meter which works in the RF field close to the antenna, e.g., a few feet, and it operates without need of batteries, a power supply, or a resonant circuit.

In use, start a few feet from the feed and come closer to the feed.

The circuit uses just 4 parts plus a radome. The most important part is the log-periodic antenna (LPD) imprinted on a piece of PCB and available in the USA and in England. The frequency range is 2-11GHz. It is pictured above.

The circuit is so very simple: the centre conductor of a piece of semi-rigid .085 coaxial line connects to a diode with output to a 25 (or so) micro-amp meter. A capacitor of about .001uF is connected between the 2 terminals of the meter and the outer conductor of the cable is connected to the other terminal of the meter. Meters of 50 or 100 microamps should be acceptable but less sensitive. A simple radome for protection against poor usage of this sniffer is an old plastic peanut can which I emptied myself.

This is hard but essential work … if it is too tough for you, you may mail the peanut cans to me for emptying , HI! There is no attenuation according to my test on my 10GHz beacon where the same radome covers the slot antenna.

The sniffer is so simple to operate: start in front of your parabola at a distance of about 15 feet if you have some power (like 40 watts) or at a few feet with QRP and meander slowly towards your antenna with the polarization like your RF source. With 40 watts, I tend to back into the dish as I can avoid looking into the feed. You will know by meter movement when your signal is strong enough or you will have a clear sign that some work is needed on your MW gear!

There are two sources for the LPD :
The UK Microwave Group and WA5VJB.
(email him at wa5vjb@flash.net)

Using 24-28VDC Coaxial Relays with 12V Supplies

Hans ON4CDU

There is always interest concerning relay power supplies.

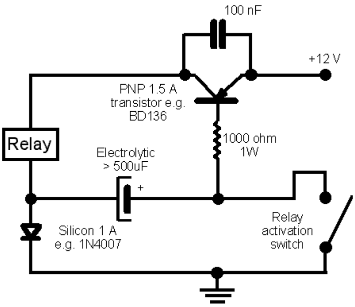
Here's a circuit based on the fact that all relays, once activated, need a much lower voltage to remain active.

The circuit originates from one by Tony, I0JX.

When the PTT switch is closed, the relay is activated by almost 24 volts, which is the sum of the PSU voltage and the charged electrolytic capacitor.

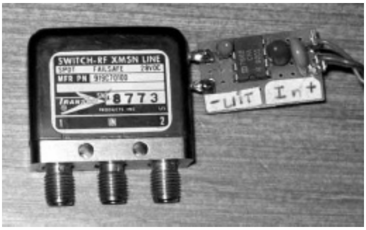
As the capacitor discharges, the voltage across the relay falls to the 12 volt level which is sufficient to hold the relay over.

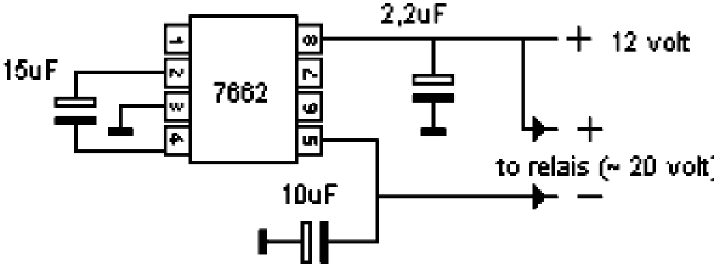
You can adjust the values of **R** and **C** to suit your relay.



Here's a simple one based upon an IC7662 chip.(see photo right:)

With just four components, it is easily assembled and can be attached directly to the relay as shown on the left in the photo. The circuit is shown below.





More Ideas for Home Made Microwave Dummy Loads

Richard T. Nadle, K2RIW

INTRODUCTION

In 1999 I submitted an article to the Microwave Reflector on the subject of the 10GHz Circular Waveguide experiments that were performed by the Ten-X Microwave Group (from Long Island, New York) using 3/4 inch copper plumbing tubing. One of the components we had constructed multiple times were Circular WG dummy loads by machining a piece of 3/4 inch wooden broomstick handle and placing it inside a piece of the 3/4" copper tubing. This made a high performance "slow absorber". If the point on the broom stick handle was sharp enough, the reflection was less than -35 dB; that's a VSWR of 1.04:1

FANCY MATERIALS REQUIRED? - Neophyte microwavers have been led to believe that very special materials are required to construct high quality WG dummy loads and attenuators. This is somewhat true only if the performance must be obtained in the smallest possible package and be maintained over all environmental conditions and the recommended frequencies for that particular WG. However, we microwavers are usually not that fussy! If a Home-Brew WG attenuator is 2" long, compared to the commercial and more expensive product that's 1" long, and if the attenuation has a slight slant with respect to frequency, I doubt the crafty Microwaver would mind. He will simply calibrate it versus frequency.

With those concepts in mind, you will soon realize that many of the inexpensive materials at your disposal can be used to construct rather high quality fixed attenuators, variable attenuators and dummy loads.

Paul Wade, W1GHZ used bicycle tyre material as a microwave absorber (see earlier pages ...editor). I believe that the small VSWR that Paul experienced would almost completely disappear if his material had been cut into a tapered (wedge-like) shape. We used wooden dowels and broomstick handles and achieved Home Brewed dummy load VSWRs as low as could be measured, once we used a proper taper.

PAD CONSTRUCTION -- Although we didn't perform the experiment, we speculated that high performance WG fixed attenuators could be constructed by placing a wooden "bullet" in the WG that had been sharpened on both ends. This technique should work equally well on Circular WG and Rectangular WG. The attenuation can be adjusted by changing the length, width and placement of the wooden "bullet", or by placing a number of bullets in the WG. If the desired attenuation is too large, when using a reasonable-sized wooden bullet, there are at least two alternate approaches:

1. **The bullet doesn't have to symmetrically fill the WG.**

A small diameter, sharp-pointed piece of wood (a dowel, or a sheet of wood) that is placed in one of the WG corners will perform admirably. The exact cross section of the attenuator doesn't matter. As long as it has a taper that is slow enough in cross sectional change per wavelength, than there will be very little reflected energy (VSWR). By moving the piece of wood from a WG corner, toward the centre of the WG, the insertion loss will increase.

This technique can be used as an attenuation Fine Tuner. Again, with proper taper, the VSWR will not change appreciably as the absorber is moved toward the WG centre.

2. **A full-sized WG low loss bullet could be constructed from Balsa wood.**

This low density material will have a much slower attenuation constant in dB per inch.

PAD CONSISTENCY

The Dissipation Factor (or Loss Tangent) that is caused by the wooden bullet is strongly affected by the moisture content of the wood. Therefore, I recommend painting the wooden bullets with a weather-proofing paint so as to maintain their moisture content (loss consistency). Depending on the Dissipation Factor of the paint that is chosen, I suspect there will be a slight loss increase after the bullet is painted. Don't judge the final insertion loss until the paint has dried. There probably are some low Dissipation Factor paints, such as lacquer, which will have very little impact on the bullet's additional loss.

MORE PAD & ATTENUATOR DATA

Concerning WG components, here's THE BOOK: George Southworth, "Principles and Applications of Waveguide Transmission", D. Van Nostrand Co., 1950; 689 pages (an oldie but goody).

Sometimes this book is available from used book web sites; it's a good buy. It contains some of the best PICTURES of how rectangular and circular WG really works with lots of performance curves (you won't need the math to understand the pictures (pages 166 & 169), it's almost an animation) -- amazing stuff for 1950.

On pages 269 to 276 you'll find pictures of linear, binomial, Gaussian, and exponential WG impedance stepping functions for broadband impedance matching, 14 designs for dummy loads (pages 368 to 371), about 25 attenuator designs.

Page 121 (A & B) has pictures are 21 of the circular WG modes (with the relative sizes of pipe shown, same frequency) made with an "RF absorbing camera".

The book shows some great transition devices, hybrids, mode killing devices & devices for launching higher modes (pages 354 to 362), round WG components (pages 269, 327 & 328), circular guide fin line (page 133), a great section explaining choke flanges (page 201), a circular pipe polarization rotating device that's "home brewable" (page 207), the shapes of circular and rectangular WG (of constant periphery) that give minimum loss (page 193 … the popular ones are not optimum), "skeleton WG" (page 175), about 15 kinds of WG irises (page 246 & 255), circular WG filters (page 307), the Qualcomm duplexing filter explained (page 309), rotary vane phase shifter (page 333), rotary vane attenuators defined (page 375), a way of designing a variable conductance dissipative film (page 377), 33 pages of horn data (only portions have appeared in other WG or antenna books), 8 kinds of "backfire" feeds including the Cutler (pages 448 to 454), eight types of WG slot antennas (pages 425 & 430), five kinds of corner reflectors, waveguide lens antennas, some TWT and magnetron info, etc. The picture on page 186 shows me how I could make S-band WG out of rain gutter down spout tubing. Let a Microwaver stand in a good hardware store with that book in hand and I think he'll get some great and crafty microwave ideas.

ARTICLE REPEAT (IN PART) -- Below I have repeated two sections of the 1999 article. The numbers refer to the section numbers of the original article; there were 14 sections.

Two sections of the 1999 K2RIW article entitled "Circular WG Frequencies, More Accuracy, More Experiment Data"

7. DUMMY LOADS

In circular WG are quite easy to construct. Simply sharpen a 3/4" broom stick handle and force it into the 3/4" copper pipe. About 3" of taper and 2" of non-taper is FB. The usual moisture in the wood makes a great "slow absorber", which makes it more forgiving of errors. The main difference between a -35 dB S11 dummy load (VSWR = 1.04, [sharp tip]) and a -20 dB S11 (VSWR = 1.22) seems to be how sharp the point was at the tip of the broom stick handle and was the taper too abrupt (too short). There may be some variations caused by knots in the wood, but we didn't seem to have that problem. The completed circular WG dummy load consists of a ~ 7" piece of 3/4" pipe with the tapered broom stick handle in it plus a copper pipe coupler at the open end. Some of the broom stick absorber can stick out the pipe far end, if you prefer. It is easy to place this load on any other piece of circular WG, while running component tests. These pipe couplers really are "sexless" connectors. For experienced rectangular WG users, it will feel strange to make connections in 2 seconds and not worry about screwing down the flanges to get a good VSWR.

9. PADS

We never did this, but it would be easy to design circular WG fixed attenuators by decreasing the length of broom stick absorber and tapering both ends to have a good impedance match from either direction. In this case I would recommend painting the absorber to keep the moisture content (absorption) constant.

If it is found that the loss is too great for a convenient length of tapered wood absorber, consider making the absorber out of six "splines" by using thin sheets of wood, or out of balsa wood. These low density materials (with tapered ends) will allow a lower insertion loss to be constructed from a longer length of wood absorber. Also, the slower loss characteristic will cause a lower VSWR for a particular taper rate.

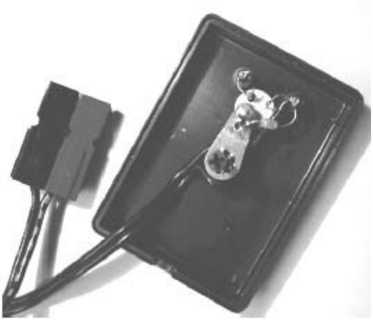
A Voltage Standard for the Ham shack

Paul Wade W1GHZ ©2004

Recently, while tuning up a new project, it seemed like the knob on the power supply turned awfully far just to reach 12 volts — then I noticed the "low battery" indicator on the digital multimeter. The voltage was already above 16 volts. Fortunately, that project has an internal regulator so no damage was done. However, it was annoying — this digital multimeter replaced one whose flaky range switch caused similar problems.

At times, I've pulled out the old Simpson VOM - the red needle is reliable but not as precise.

A few nights later, I found a bag of parts buried on the workbench — probably purchases from a hamfest last summer. Among the gems was an IC marked AD581L. A quick search at www.analog.com showed it to be a precision voltage reference, laser-trimmed to exactly 10 volts. Just the ticket for untrustworthy digital meters. The AD581 has only three terminals: an input of 12 volts or higher, an output of 10 volts, and ground. No external components required.



I dug up a small ABS plastic box, a couple of pin jacks, and a cord with an Anderson Powerpole connector. Two holes and three solder joints completed the assembly. The photo shows the complex assembly — I added a bypass capacitor, just because. A schematic is hardly necessary.

Time to spark it up — the output was 10.01 volts on the digital multimeter. Pretty good, but the L suffix is specified to be within 5 millivolts, so I tried a lab-grade meter, and read 10.003 volts — really good. The data sheet talks about aging for 200 hours to stabilize, so I ran it for a week and measured again: 10.0031 volts. Is this accuracy necessary? Probably not, most of the time.

I tried several digital multimeters of various age and quality, and obtained the following readings:
10.10, 10.03, 9.96, and 9.98 Volts, a range of 70 millivolts.

None is off by more than 0.4%, not bad considering that these were cheap meters rated for 1% accuracy when new, and never calibrated since. 1% accuracy is fine for most measurements, but some things, like battery testing and charging, require finer resolution. We've learned the hard way that good, fully-charged batteries are essential for successful portable operation.

Comparison to an accurate standard is a good way to get more accurate results.

This is a handy little gadget that you can build, even if your homebrewing skills aren't quite up to tiny surface-mount microwave components. The AD581 is readily available from

www.analog.com,

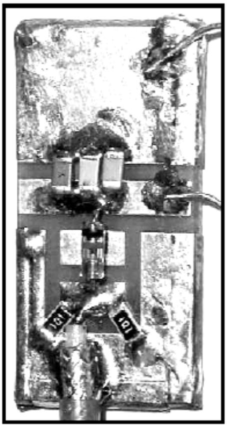
or an equivalent part, the LT1031 from **www.linear.com** (see application notes AN82 and AN42); both are available from Digikey. In either case, you can pay for as much accuracy as you need. Other voltages are also available. Many are lower noise than normal voltage regulators, so they might be good voltage references for low-noise oscillators. Some of these parts may be available as surplus — mine apparently was — so let us know if you find a good source.

A Simple but Handy Diode Detector

From the F.U.N New Zealand VHF/SHF/UHF Newsletter

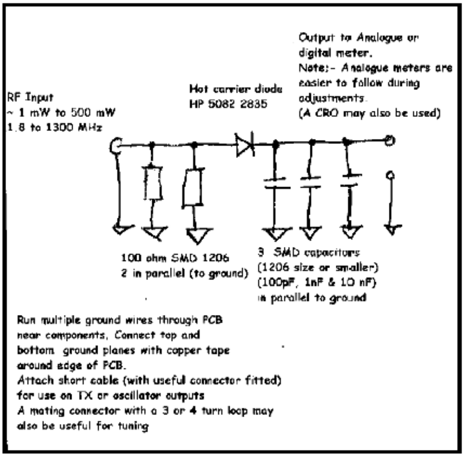
Component list for diode detector

Hot carrier Schottky diode HP5082-2835 (or similar)
100 ohm SMD 1206 resistors. (2 off)
100 pF SMD 1206
1nF SMD 1206
10nF SMD 1206
Offcut of double sided PCB 1.6mm
(Run wires through PCB near the resistors and capacitors



Parts are commonly available via the VHF Groups/ RS/ Farnell/ scrap pcbs and other sources. Contact the scribe if you have difficulty locating suitable components.
The detector can be connected to the output of an oscillator/TX via a short coax cable and connector or used with a multi-turn loop to detect oscillator/ multiplier operation. The detector works from ~1 mW to 500 mW (maximum dissipation of the 1206 resistors. The detector may be used from 1.8 MHz up to 1300 MHz. Adding a further capacitor will lower the LF limit. If required a chart can be drawn to convert the volts to power in mW, using the formula.
P (in mW) = 1000 × $[(V + 0.25)^2/ 100]$

Thank you those who have sent information for the column
Input for the column may be sent to Kevin ZL1UJG at rfman@xtra.co.nz



Output to Analogue or digital meter.
Note:- Analogue meters are easier to follow during adjustments.
(A CRO may also be used)

Run multiple ground wires through PCB near components, Connect top and bottom ground planes with copper tape around edge of PCB.
Attach short cable (with useful connector fitted) for use on TX or oscillator outputs
A mating connector with a 3 or 4 turn loop may also be useful for tuning

A low-cost high-performance RF power sensor

by Herbert Dingfelder, DL5NEG

What is it good for?

A lot of radio amateurs are afraid of building gear for GHz frequencies because they fear that they have no chance to measure at these frequencies. Well, the most interesting parameters of a signal are frequency and amplitude. Reasonably priced frequency counters up to 3GHz are widely available now but RF power meters are still quite expensive. This circuit transforms a RF signal into a DC voltage level which can easily be measured with a low-cost multimeter. Using the graph shown later in this article, you can then determine the RF signal level.

How does it work?

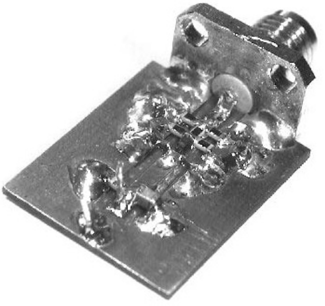
The 5 resistors at the input form a combined 3dB resistive pad and RF termination. The RF diode (BAT62-02W) rectifies the signal, the capacitors on the DC side form a broadband short circuit for RF. Having the different capacitance values in the arrangement shown (the pF very close to the diode) is very important for a flat frequency response.

The **component layout** shown on the next page is for scaling. One grid square is one millimetre.

The ground vias are very important for a constant frequency response. The more the better.

I have marked my vias in the drawing to give you an idea where the vias are most important.

I have used silver coated copper wire with 1mm diameter, the thicker the better. I have used FR4 PCB material with a thickness of only 0.8mm to keep the ground vias short. You can use 1.5mm material as well but the stripline has to be 2.5mm wide then.



RF input is the SMA connector on the left side of the circuit board. The rectified DC voltage is present on the stripline right to the diode. Just measure the voltage across the markings shown as GND and DC out. See below for finding out how much RF input power causes a certain DC level at the output.

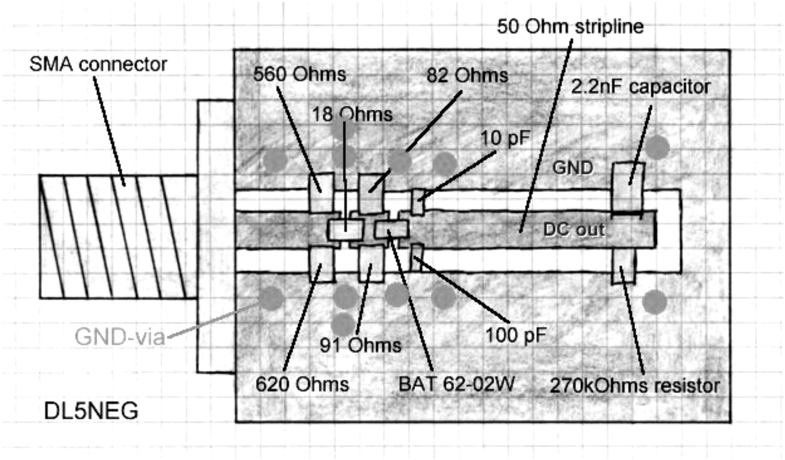
How can I convert the DC level to the RF power level?

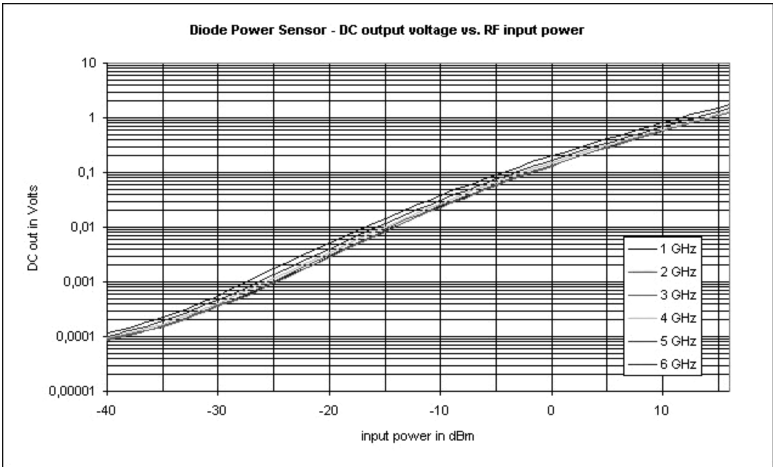
I have built a number of these circuits and the variation from one sample to another is very small.

So if you built such a circuit for yourself and stick to the component mounting drawing above, you can use my measured values. You can even use BAT62 diodes in different packages for frequencies up to 2.5GHz. Above 2.5GHz it makes sense to use the very small package that I have used to keep the self-resonance frequency high.

In the graph below you can see the DC output voltage that the circuit gives at certain input power levels at certain frequencies. ***(This graph can be seen in full colour on DL5NEG's website)***

PCB layout


SMA connector
560 Ohms
82 Ohms
50 Ohm stripline
2.2nF capacitor
18 Ohms
10 pF
GND
DC out
GND-via
91 Ohms
100 pF
620 Ohms
BAT 62-02W
270kOhms resistor
DL5NEG


Diode Power Sensor - DC output voltage vs. RF input power
DC out in Volts
10
1
0,1
0,01
0,001
0,0001
0,00001
-40
-30
-20
-10
0
10
input power in dBm
1 GHz
2 GHz
3 GHz
4 GHz
5 GHz
6 GHz

As you can see the DC output voltage is practically independent of frequency up to 3GHz. Above 3GHz there is a slight frequency dependency but, using the graph above, you can still determine the RF power with high accuracy. If you feel that the graph does not give you precise enough information you can download the complete table from my website as text file.
Go to **www.dl5neg.de**
Any questions? Just don't hesitate to contact me by email and I will be glad to provide you with answers.

Email: homepage.feedback@dl5neg.de

This diode sensor by itself will be interesting enough to many readers but I would also like to draw your attention to the sub-page on my website that describes a full-blown handheld power meter, using the aforesaid diode sensor (amongst others). It provides the radio amateur with a fully featured power meter for very little money.
Due to the fact that the measured sensor tables are put into the software (with interpolation within the measured 1dB steps), the reading is extremely precise, usually well below 0.5dB absolute error. Details can be found at:
http://www.dl5neg.de/powermeter/powermeter.html
I have another page that gives an overview on the working principles and the pros and cons of diode, thermal and log amp sensor and
you might feel it’s worth visiting at:
http://www.dl5neg.de/powermeter/theory/powersensors.html

Tuned Audio Level Meter for RF Measurements

Roger Blackwell, G4PMK

At Round Table meetings, great interest is always generated by the antenna gain measurements. Central to these measurements is the use of tuned audio level meters such as the HP415 for measuring signal levels. I published a design for a home-brew 1 kHz audio level meter some time ago [1], which does the same job and can be used both for antenna range work and many other microwave measurements.

So here is a revised and updated version. It's a tuned audio amplifier for use with 1 kHz modulated sources (both commercial signal generators and home-brewed sources) and an external diode detector. It has a dynamic range of over 50 dB. I'll cover the companion modulator and some of the uses of the meter in a future article.

The design consists of an input section, three amplifier-filter sections and a precision rectifier and meter buffer. The input section includes provision of a small amount of forward bias of either polarity, which will improve the square-law response of most common detectors. The audio driver transformer provides a suitable AC load for the detector, allows injection of this bias to the detector and provides a modest voltage step-up. The 10 dB step attenuator is split across the signal chain in the interests of best dynamic range. The circuit runs from two alkaline PP3 batteries which allows portability. An audio output is always useful was when making antenna measurements, so this design has an audio output, intended for use with one of the ubiquitous battery-powered PC/personal stereo active speakers.

This output provides about 100 mV pk-pk into a 5k load when the meter reads 100% FSD. The instrument should be constructed in a metal case using good audio construction techniques, including a single common connection for the case and circuit at the input socket, which ought to be a BNC or similar type. You can build the circuit on matrix board if you wish (as was one of the prototypes), although a PCB artwork should be available via the Internet [2] by the time you read this.

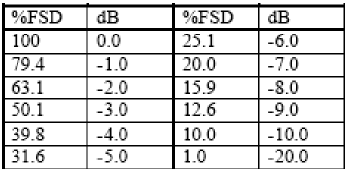
All connections to the input circuit, step attenuator and set level control should use screened leads.

Some of the components require a bit of care in selection. The 10nF capacitors in the filter stages (2 per stage, a total of 6) should be ±5% or better tolerance types. Resistors should be ± 2% or better, metal film or similar. The three presets are used to trim the filter centre frequencies and can be small cermet types. The 10 dB step attenuation switch should be a two-wafer type so as to maximise the isolation between sections. The bias switch is a single pole centre-off type. The input transformer used in the prototypes was the venerable LT44 transistor inter-stage transformer, still available from Maplin (HX82D) as are other suitable transformers (PX79L or PX80B). The PCB layout and prototypes used an LT44.

It makes sense to get the best and largest meter you can find for this project, since you will be taking measurements directly off the meter scale. The 4u7 electrolytic across the meter terminals was added to damp out jitter due to noise; depending on your meter movement and personal preference you may find it unnecessary.

The meter will need re-calibrating in

dB according to the following table:



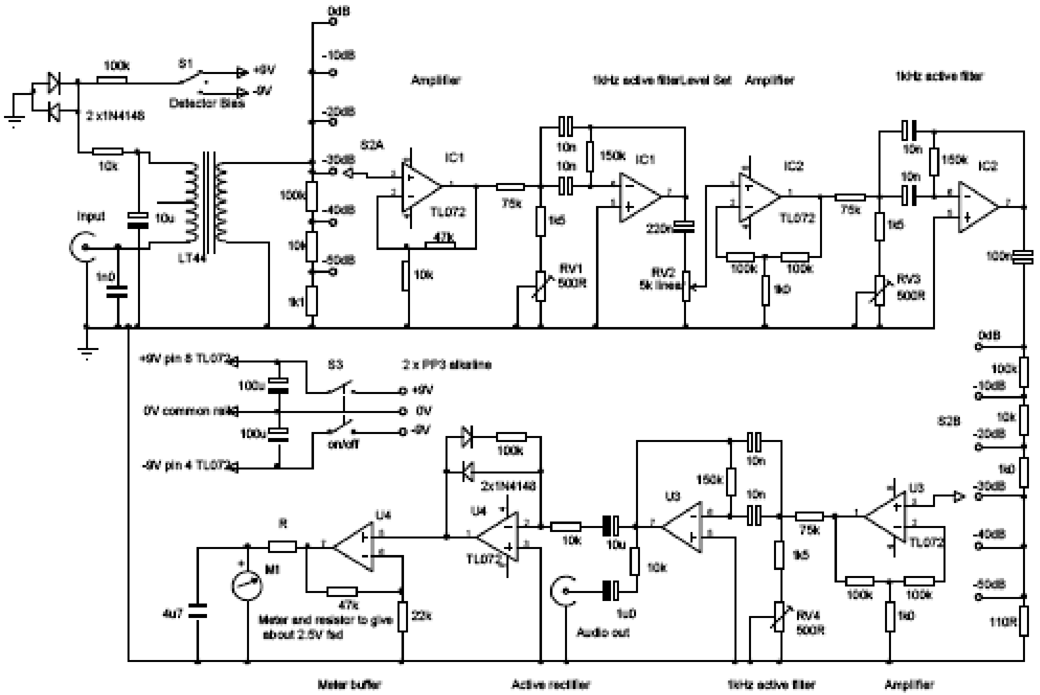
%FSD	dB	%FSD	dB
100	0.0	25.1	-6.0
79.4	-1.0	20.0	-7.0
63.1	-2.0	15.9	-8.0
50.1	-3.0	12.6	-9.0
39.8	-4.0	10.0	-10.0
31.6	-5.0	1.0	-20.0

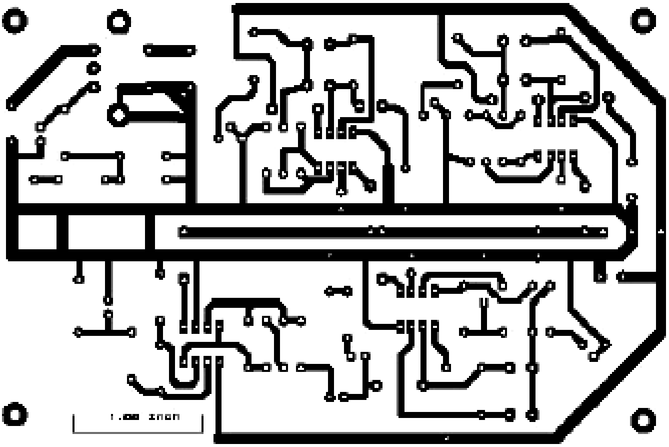
If you leave the original scale in place, then any reading can be converted to dB with the formula

$\mathrm{dB}=10.\log_{10}(100\%\mathrm{FSD}/100)$

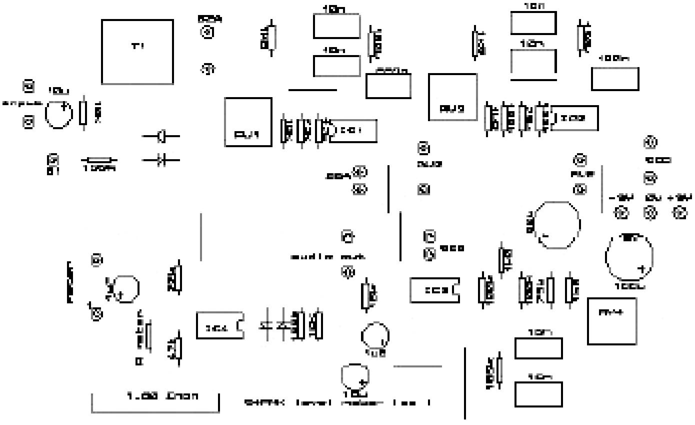
which is easy enough with a calculator.
Setting up the meter is quite simple. Inject a low-level (few mV) 1 kHz tone into the input, and adjust the attenuator and level set control for an on-scale reading. Ideally do this with the 1 kHz modulating source you plan to use. Now peak each of the three filters in turn by adjusting the 500R presets. Check that with a short-circuited input residual circuit noise is about 20% FSD on the 0dB attenuator range with the level control set to about 50%.
That's all there is to it!

The prototype sensitivity was such that about 100 mV RMS gave 50% FSD on the -50 dB setting with the set level control at mid range.
[1] Receiver Measurements, Ch 12, The VHF/UHF DX Book, edited by G3SEK,
RSGB.


Amplifier
1kHz active filterLevel Set
Amplifier
1kHz active filter
100k
S1
+9V
-9V
Detector Bias
2 x1N4148
10k
Input
10u
1n0
LT44
0dB
-10dB
-20dB
-30dB
-40dB
-50dB
S2A
100k
10k
1k1
IC1
TL072
47k
10k
75k
10n
10n
150k
IC1
1k5
RV1
500R
220n
RV2
5k linear
IC2
TL072
100k
100k
1k0
75k
10n
10n
150k
IC2
1k5
RV3
500R
100n
0dB
100k
-10dB
10k
S2B
-20dB
1k0
-30dB
-40dB
-50dB
110R
+9V pin 8 TL072
0V common rail
-9V pin 4 TL072
100u
100u
S3
on/off
2 x PP3 alkaline
+9V
0V
-9V
100k
2x1N4148
U4
TL072
10k
10u
10k
1u0
Audio out
U3
150k
10n
10n
75k
1k5
RV4
500R
U3
TL072
100k
100k
1k0
R
U4
M1
47k
22k
4u7
Meter and resistor to give about 2.5V fsd
Meter buffer
Active rectifier
1kHz active filter
Amplifier



Track Layout



Component Layout

Using the Tuned Audio Level Meter

Accessories

To use the level meter, you need to provide an RF source, amplitude modulated at the 1 kHz frequency to which the level meter is tuned. If you have a signal generator or QRP Tx capable of amplitude modulation, that will do nicely; if not then it's a simple job to provide an external modulator for whatever RF source you use.

Fig 1 shows a low-level (mW) modulator based on a PIN diode and a simple oscillator. By using appropriate construction techniques, SMD style components and sensible values for the RF chokes and bypass capacitors, this can be made to work up 2 GHz or so. Make the audio frequency adjustment easily accessible so it can be trimmed for best response with the level meter. The 2-pole centre-off switch allows selection of both modulated and unmodulated signals without disconnecting the modulator from the source.

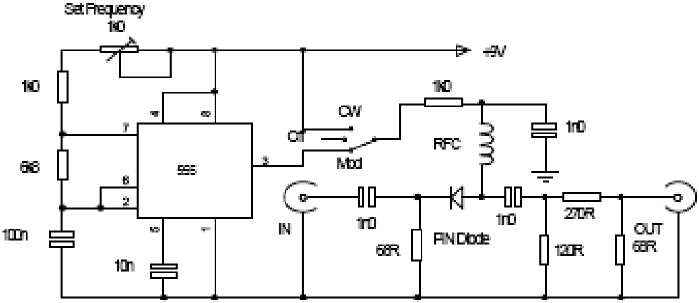


Fig 1: 1kHz amplitude modulation accessory

For higher frequencies, it may be possible to utilise a commercial semiconductor switch; I have used the Mini-Circuits ZFSW series to 5 GHz or so, and there are occasionally waveguide PIN modulators or switches available as surplus. Actually, I have a confession to make; I use a klystron source for 10GHz, modulated by feeding 25 V or so via a capacitor to the reflector. Sufficient AM is produced to render this source quite suitable for measurement applications. A Gunn oscillator could no doubt be modulated in a similar way.

The other item needed to complete your measurement outfit is a detector. One can often find suitable detectors on the surplus market - I'm thinking of those with an N-type plug on one end and BNC socket on t he top, or one of those SMA m/f types, both look something like a coaxial attenuator. The diode type and polarity will determine what setting of the bias switch to use. Remember that this is a series diode, so the device you are connecting it to should provide a DC return for the bias current.

Often a -3dB or larger attenuator between detector and source is a good choice and serves as well as a normalising impedance. If you cannot come across a surplus detector, then make your own in the body of a N-plug, with the diode connected between the centre pin and a feed through capacitor fitted on a disk of brass soldered on the top of the clamping nut.

If you are lucky enough to find one

at a reasonable price, a *slotted line* should have a detector fitted on the sliding carriage, and will enable you to make vswr measurements of components very easily.

Antenna Measurements

The interest shown in making antenna measurements using tuned audio level meters was what originally prompted these articles. I'll refer you to some references [2,3,4] for a full treatment, however here are some of the basics. This incarnation of the level meter has a separate audio output, designed for use with an external battery-powered amplifier/speaker. This is useful, not only to hear what's Really Going On, but also to aid alignment of the antennas; if you are changing antennas for comparison purposes, it's particularly important that both are aligned for maximum forward gain rather than just visually.

Gain Measurement

For accurate gain measurement - or even accurate comparisons between different antennas - it is essential that the antenna under test is located entirely within a three dimensional region of uniform field strength with a flat phase front. You will need a *source antenna* to create this RF field, and if you want to measure absolute gain you will also need a *reference antenna* of known gain. If you just want to compare two antennas, the reference antenna isn't required. The reference antenna could be one that's been measured elsewhere, or a standard gain horn, which can be home-made from published data.

For the purposes of this description, the source antenna is treated as the transmitter, and the antenna under test as the receiver. This is usually the most convenient setup, but the transmitter and receiver can be interchanged.

To create the uniform field mentioned earlier, the source must be in the *far-field* of the dish under test, and vice-versa. That means a distance greater than:

$$2D^2 / \lambda$$

between the two antennas, where D is the diameter of the dish, or the greatest dimension across the aperture for other antennas. If you are concerned with Yagi antennas, you can calculate the equivalent aperture (a.k.a. capture area):

$$D_{EQ} = G\lambda^2/4\pi$$

where G is the gain over isotropic (in linear, not dB terms), ? is wavelength. D_{EQ} can then be substituted for D in the equation above.

For example, a 60 cm dish will have to be at least 24m away from the source at 10 GHz.

The ideal antenna range is floating in free space, and several ingenious range geometries have been devised to avoid the effects of ground reflection - if you can measure across a deep canyon or between two high buildings, then do it! However, for most amateur measurements we have to use a level range and cope with the effects of ground reflections somehow. In most range geometries, both of your antennas need to be well clear of the ground and other reflecting objects. In general try and choose a level site for measurements - free of buildings, cars trees and metal objects. Ideally the surface between the two antennas should be flat, and smooth to a quarter wavelength or so - which is no doubt why measurements are often made in large empty car-parks. If possible, use vertical polarisation of the antennas as then the angle of incidence of the transmitted signal with the ground can be set to minimise reflections - this is known as the Brewster angle and is about 15° for soil surfaces. The spacing between the source and receiving antennas needs to be 7.5 times the source antenna height to achieve this Brewster angle. However, you also need to meet the far-field criterion mentioned earlier, and this may require impractical heights for the receiving antenna. In any case, it should be possible to check that you have a suitable set-up, by investigating the signal strength across the plane of the receiving antenna, using a dipole-on-a-stick probing antenna and the level meter. The signal strength should remain constant (within a dB or so) across the whole volume into which you are going to place your test antennas.

Dish Focussing

This is pretty much the same as gain measurement, but obviously doesn't require a reference antenna. You'll still need a source and a transmitting antenna though.

Having the detector/level meter attached to the dish to be adjusted will be more convenient. The level meter is also well suited to setting up circularly polarised feedhorns.

Gains and losses in components

Because of the dynamic range possible with the level meter, it is a very useful tool for measuring gains and losses - in cables, amplifiers or just about anything else. Of course, it doesn't matter if frequency conversion takes place, as long as there is an output at some frequency that can be fed to the detector. It's important to check that you are keeping within the linear characteristics of any active devices. Filters are a particular favourite application of mine - a band-pass filter response for example can be quickly plotted with a signal generator and the level meter. Very handy if you don't have access to a sweep generator or network analyser.

Using a Slotted Line for matching and impedance measurements

A slotted line is one of the most useful accessories for the level meter. These are sometimes available on the surplus market. You're more likely to come across a waveguide version, which consists of a length of guide with a slot along the top.

Some kind of movable carriage runs along the top, with a probe whose end projects through the slot. A detector is connected to the probe. These lines are also made in coaxial airline form as well. You can use them for making SWR and impedance measurements, using just the level meter and a modulated source. Here's how:

Simple VSWR measurement

VSWR is defined as the ratio of maximum and minimum voltage on the line:

VSWR = Vmax /Vmin

The square law detector used with the level meter means what is indicated on the meter is a power ratio, so:

VSWR = v (Vmax/Vmin)

First slide the carriage along for maximum indication, then use the Level Set control on the level meter to set FSD (100) on the meter. Now adjust the carriage for a minimum reading. The VSWR is then:

√ (100/reading)

If you have actually calibrated the meter in dB, then FSD will be 0dB, and the minimum reading is a 'loss' in dB , so:

VSWR = antilog$_{10}$ (Loss dB/20)

Impedance measurement

This is a bit more involved, so I'll go through it in steps.

A) Place a reference short (for waveguide, a metal plate) across the output side of the slotted line. Find the positions of TWO peaks, and hence the distance between them.
That's the guide wavelength, $\boldsymbol{\lambda_g}$.

B) Connect the device to be tested in place of the short. Move the carriage to a maximum, and set the meter for FSD. Note position of the carriage.

C) Move carriage either towards the load or towards the generator (remember which!) for minimum. Note the position of the carriage again, and read the meter scale for

VSWR as described earlier.

D) Now dig out the pad of Smith charts [5] and draw a VSWR circle on the chart.

E) Translate the distance between the two points already noted (max and min) into fractions of λ_g using the length of λ_g obtained above in step C.

From the left hand side
(R=0) of the chart, move the appropriate fraction of a wavelength around the circle— forward = towards load = clockwise; backwards = towards generator = anticlockwise.

The point on circle is the impedance in normalised terms.

If you are fortunate enough to have a 50 ohm coaxial slotted line, then the impedance is normalised to 50 ohms. If you are using a waveguide slotted line, then the waveguide impedance is:

$$377 * (a/b) * (\lambda_g/\lambda_o)$$

where λ_o is the free space wavelength, and a and b are the internal dimensions of the waveguide.

I've used this technique very successfully with a WG16 slotted line to make and adjust WG to coaxial transitions for 10GHz.

So that's a few applications of the Tuned Audio Level Meter - of course there are others. Why not write up your application for *Scatterpoint*?

[1] Scatterpoint Issue 4.
[2] Antenna Gain Measurements, Fred Brown W6HPH, QST Nov and Dec 1982.
[3] IEEE Standard Test Procedures for Antennas, ANSI/IEE 149-1979
[4] UHF Antenna Ratiometry, Richard Knadle K2RIW, QST Feb 1976.
[5] *In Practice*, RadCom March 2001.

A downloadable bitmap of the Smith chart is at
http://www.ifwtech.com/g3sek/in-prac/smith.gif

1kHz Amplitude Modulated Gunn Oscillator for RF Measurements

David Wrigley, G6GXK

A useful accessory for the previously published "Tuned Audio Level Meter" (ref. 1) is this simple 10GHz Amplitude Modulated Gunn Diode Oscillator.

Using this equipment, tests can be carried out on antennas and other parts of your 10GHz equipment, using the significant dynamic range of the tuned audio level meter.

Basics

The obvious choice for a 10GHz source is a Gunn diode – it is very low cost, simple and easy to operate. The one problem is that it isn't so easy to amplitude modulate.

The straightforward application of a modulating voltage to the Gunn diode produces lots of FM, which may well cause problems due to unexpected off frequency responses. A separate pin diode or FET modulator was the next thought, but that would require two transitions to get to SMA and back to WG and a horn. It was then that the Solfan combined Gunn diode and detector combination was considered as a possible source of AM. Such unit have been used in wideband transverters and one of these was selected for some preliminary tests. This particular unit came from a PW "Exe" WB transceiver had a piece of 2BA screwed plastic rod (knitting needle) entering the cavity as a fine frequency adjustment.

The Gunn was fired up with 7.0V and the output adjusted with the aid of a cavity wave-meter to be approximately 10368MHz. With a power measuring device (diode detector) in front of the Gunn diode assembly, the effect of passing a 5mA current through the Solfan detector diode was checked. A quite definite change in output level was seen of about 10 to 20% in amplitude. The extent of this could be peaked by adjustment of the screws around the detector diode, which project into the waveguide. Further checks with a 1KHz modulating signal were carried out. These proved that the resulting modulated Gunn output had no problem in end-stopping the Tuned Level Meter when fed from a simple diode detector and small horn. This has surely got to be the simplest, lowest cost method of getting 10GHz AM!

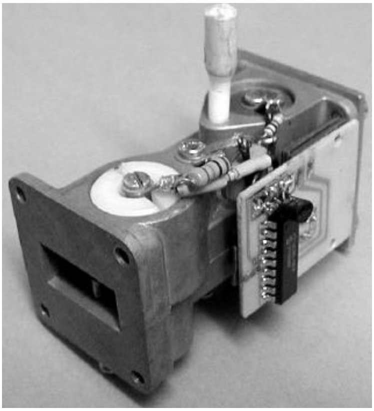
Audio Source

The next thought was to generate an accurate 1KHz signal. Past experience had shown that it is not so simple to produce a 1KHz signal with sufficient stability and accuracy to reliably pass through the narrow filters of the tuned level meter. The final solution was to use a PIC16F84 microprocessor to give precisely 1KHz . Well, within about 40ppm, and that is amply good enough.

The program is based on the fact that the instruction period of a PIC microprocessor with a 4MHz crystal is precisely 1 microsecond. Therefore the program consists of two 500us delays interspersed with toggling an output line, and then finally looping back to continuously repeat the process. The result is a precise 1KHz square wave. Since the harmonics are outside the pass-band of the tuned level meter, there is no problem. The whole PIC generator consumes about 6mA including the LM78L05 regulator. The PCB was designed to use "In Circuit Programming" so wires can be attached to the 5 pads provided and fed from the programmer socket. To program this PIC chip "in-circuit" you need to connect 5 wires to the appropriate pins of the programmer socket as detailed on the circuit in the zip file and just use the programmer as you would normally – the power is supplied from the programmer so no other supplies are necessary. Alternatively you could program the PIC before assembly or a ready programmed PIC chip can be supplied (ref. 2). A zip file of the overall circuit, the PCB layout and the hex file for programming is downloadable from our website. (ref. 3).

The PCB was stuck to the Gunn diode assembly using thick double sided adhesive tape. (as used in the automotive industry – this is so good that it will even stick on registration plates!).

The picture below shows the basic assembly.



It had been intended to supply the whole thing from a 12Volt Battery pack and the trial unit used a 7V 1.5A regulator, which was to hand. A small regulator PCB with an LM 317 will eventually be built to take over this job. The total consumption including the Gunn diode is about 100mA. And the TO220 style LM317 should have no problem handling the 0.5W dissipation without any special heat sink. Once the basic system had been built up, further trials took place. At about 25metres the level meter showed about half scale on the –10dB range using a small horn. Use with a dish would give considerably more output and therefore this was considered to be acceptable.

Some further lessons were learnt from these trials:-

1. A proper fixing was required for a camera pan and tilt tripod
2. Some mechanical protection was required for the PCB and Gunn assy.
3. A simple sight was required for the initial lining up of the transmitter. Even a small horn is quite strongly directional.
4. It is essential to make sure that the batteries in the tuned level meter are fully charged before a long measuring session.

Failure to get a sensible reading can result in all sorts of unnecessary checks and adjustments.

The first three items were dealt with by designing a sheet metal cover to support and protect the assembly. This cover was built from 0.8mm thick Zintex (zinc plated steel) sheet which was folded and MIG welded. This could equally well have been soldered with large fillets.

The cover had a simple sight (hole and cross-wires) built into it. It also had two 0.25inch x 20tpi nuts welded in to provide tripod attachment. A full sized drawing for the cover is downloadable from our website within the zip file previously mentioned. The picture on the following page shows the final assembly mounted on a camera tripod.

Conclusion

So having built up the 10GHz transmitter, this completes the equipment required to carry out comparative tests on 10GHz antennas.

These are:

1. Modulated Gunn diode source with small horn – as described here.
2. Battery and regulated power supply for the transmitter
3. Waveguide mounted diode detector assembly and small horn (would be normally fitted directly to the illuminating horn of the dish under test or via a transition to an SMA connector output of the horn)
4. AF tuned level meter to the Roger Blackwell design published in Scatterpoint issue 4 and 5.

10GHz is perhaps the most commonly used microwave band – it certainly one of the easiest to get onto. This equipment will help us to get the best from our equipment and also demonstrate when we have succeeded in improving it

References

1. Tuned Audio Level Meter for RF Measurements, Roger Blackwell G4PMK, "Scatterpoint" –issues 4 & 5.
2. Email the author: g6gxk@arrl.net
3. UK microwave group website:
   www.microwavers.org

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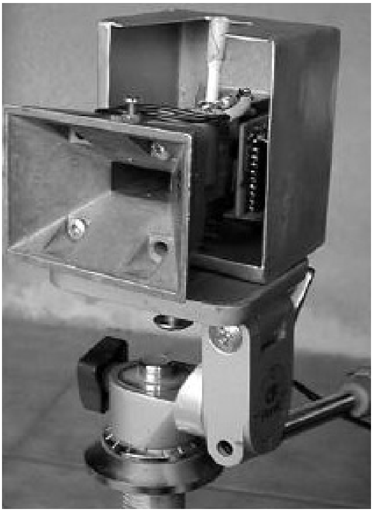
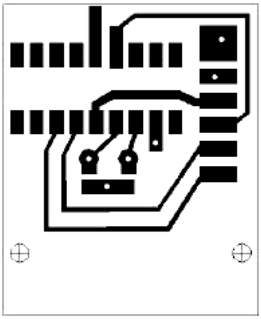
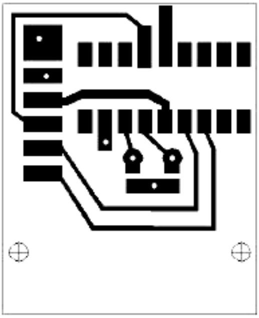
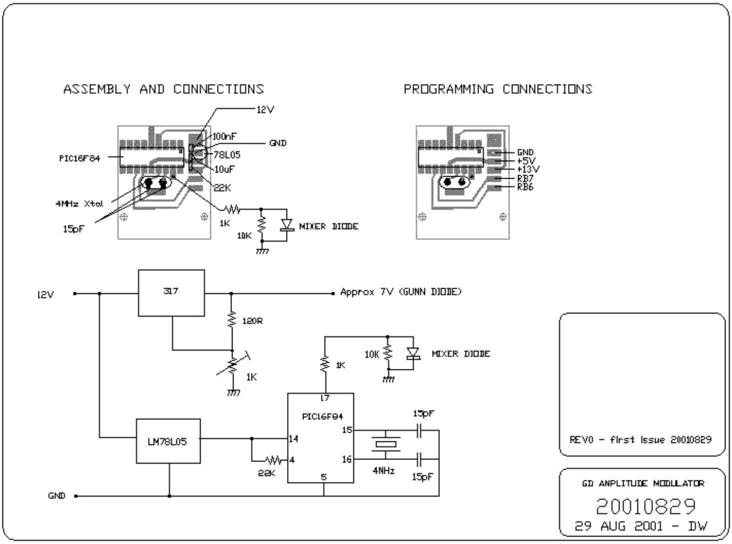




COPPER



MIRROR IMAGE
FOR IRON-ON RESIST


ASSEMBLY AND CONNECTIONS
PROGRAMMING CONNECTIONS
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100nF
GND
78L05
10uF
22K
PIC16F84
4MHz Xtal
15pF
1K
10K
MIXER DIODE
GND
+5V
+13V
RB7
RB6
12V
317
Approx 7V (GUNN DIODE)
120R
1K
10K
MIXER DIODE
1K
17
PIC16F84
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LM78L05
14
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22K
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4NHz
15pF
GND
REV0 - first issue 20010829
GD ANPLITUDE MODULATOR
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Single LED Battery Status Indicator for the 'Rover'

Paul Wade, W1GHZ © 2005

Fading batteries are an all too frequent part of rover operation. Battery voltage fades slowly and silently, usually unnoticed until something doesn't work quite right, and then the problem is still sometimes overlooked. A simple warning device that doesn't distract from normal operation would help.

Battery status is easily monitored by the battery voltage – a "12 volt" battery, like a car battery, delivers around 13 volts when fully charged, dropping to below 12 volts when nearly discharged. A rover operator could check this periodically, but there are periods of attention overload: simultaneously digging out weak signals, maintaining liaison, doing PR with onlookers, and keeping antennas from blowing over in the wind can fully occupy most of us. So, while we could easily read a DVM, or just push the right button sequence on an FT-817 to measure the voltage, it may get neglected until something doesn't seem right.

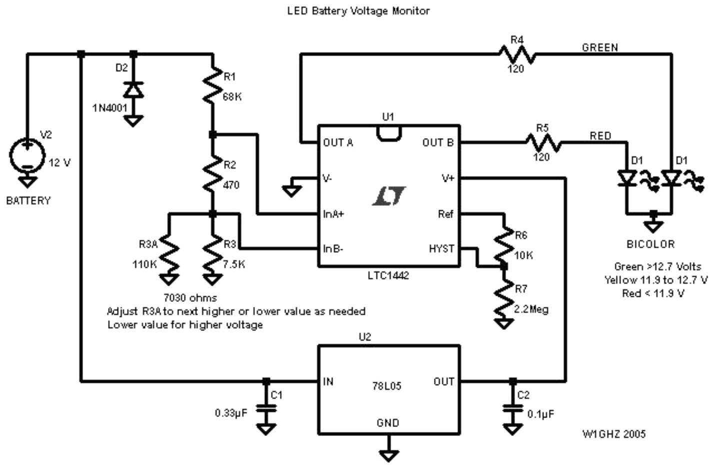
A simple indicator, like an LED, is much easier to notice without being a distraction. It occurred to me that a multi-colour LED could be used as a status indicator. The most common combination is red and green; obviously, green for good and red for bad. When both colours are lit, the output is yellow, which could be a warning state. Thus the simple display would be green if the battery is good, yellow when the battery is getting low, and red when it is dying.

Now we need a simple circuit to detect the voltage and drive the LED. Integrated circuit comparators are cheap, but the difference between a good battery and a dead one isn't large, so the voltage must be reasonably accurate. The comparison voltage, reasonably accurate over time and temperature, may be obtained from an integrated circuit voltage reference at a modest price.

A comparator has two output states: high when the "+" input is at a higher voltage than the "—" input, and low when the "+" input is at a lower voltage. To drive the bi-colour LED, one comparator should turn off the green light when battery is nearly dead and another comparator should turn off the red light when the battery is near full charge. In between, both lights are on, so the display will be yellow.

I have some LM393 comparators, the cheapest one available, in the junk box. However, all the available bi-colour LEDs have a common cathode lead and separate anodes, so the open collector outputs of the LM393 would not drive the LED without external transistors – too many parts for a simple circuit. A bit of research located a dual comparator with a built-in voltage reference, the LTC1441 (**www.linear.com**), with a price comparable to a voltage reference alone. The only drawback was that this comparator will not operate at 12 volts, so a 5-volt three terminal regulator is necessary to reduce the voltage. The complete circuit is shown in the schematic, **Figure 1** on the next page.

The voltage divider R1, R2, and R3 sets the voltage trip points for the comparators in U1, so that the red light goes out at voltages above about 12.7 volts and the green light goes on at voltages above 11.8 volts.
Since I used ordinary 5% resistors, the necessary adjustment for tolerances is



provided by R3a. If the trip points are too low, R3 is decreased by replacing the nominal value of R3a, 110K, with the next lower standard value, 100K, or 91K if necessary. If the trip points are too high, R3a is increased to 120K or 130K.

Once the adjustment is made, the resistors will probably drift and age together to maintain the desired ratio.

When I bread boarded this circuit, it appeared to work fine, with the colour changing as planned.

However, when the voltage was set exactly at the trip point, thermal noise switched the comparator on and off randomly at a rate invisible to the eye but visible on an oscilloscope. The comparators are too sensitive! The solution is to add some hysteresis, so that there is a slight difference between the on and off trip points. The LTC1441 has a pin for adding hysteresis, by adding R6 and R7. The values shown provide about 0.1 volts difference between the on and off voltages.

My breadboard didn't look robust enough for rover use, and I had a bit of spare area on a PCB for another project, so I made a quick layout of Figure 1 and added it to the board, piggybacking on a $59 Miniboard with ExpressPCB **www.expresspcb.com**

Figure 2 is the PCB layout for use with the free ExpressPCB software – the plain version I built and two others with PowerPole connectors on the board, to monitor voltage as it enters a box. The file is available at

www.w1ghz.org

A photo of the complete battery status indicator is shown in **Figure 3.** Small and simple, with one LED, two ICs, eight resistors, and two capacitors. Total parts cost, new from DigiKey (**www.digikey.com**), is about $3. I found three choices for bi-colour LEDs in the catalogue: the 160-1057-ND had the best three colour combination, the 67-1124-ND was pretty good, but the 160-1715-ND had a very greenish colour when both are lit so the yellow state is harder to discern. Any of the three LEDs run about $0.30, so choosing the best colour doesn't cost extra. I included the unit in Figure 3 inside my Battery-Sharing Switch, so I now have one box to control my batteries with just one light to keep an eye on.

Fig. 2

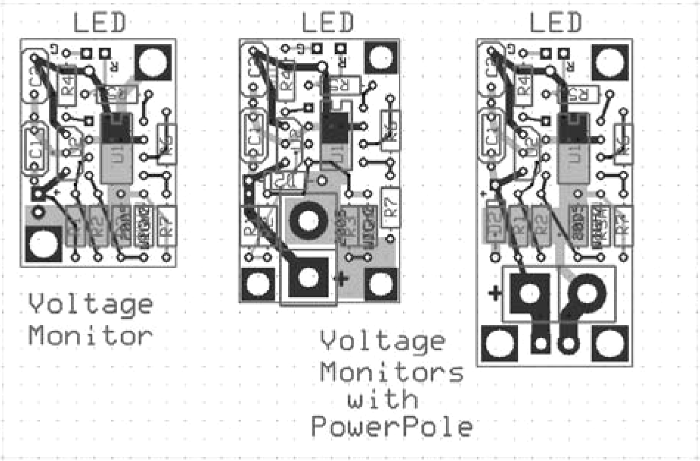
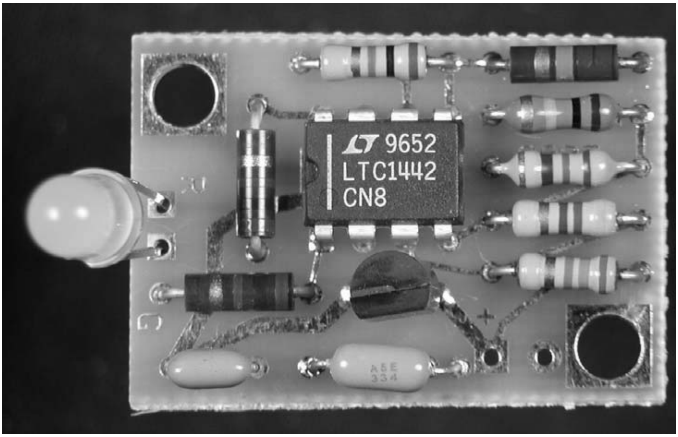
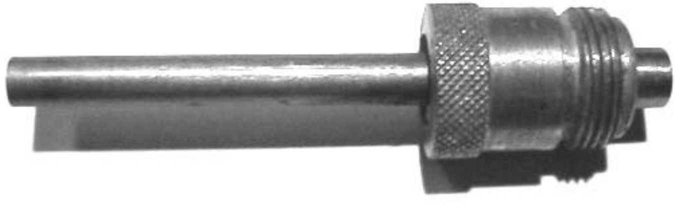

LED
LED
LED
Voltage
Monitor
Voltage
Monitors
with
PowerPole
C1
C2
R1
R2
R3
R4
R5
R6
R7
U1
+

Fig. 3


9652
LTC1442
CN8
R
G
+

Circular Waveguide to N type Adapter for 47GHz Power Measurements

Kerry Banke, N6IZW



As I started getting my feet wet with experiments on 47GHz and above, I was looking for a way to roughly measure power. I have an HP436A, Gigatronics 8540C, and PMI 1018B, all with heads rated to 18GHz .

The questions were would any of these heads respond at all to 47GHz, how would I couple 47GHz to the power head and how could I do a rough calibration? The calibration turned out to be fairly easy as Will Jensby, W0EOM, loaned me a multiplier module for which he had measured the power output.

The experiments that myself and Don Nelson, N0UGY, had been performing mostly used 0.188 inch ID hobby tubing as circular waveguide so I needed a transition to the type N connectors on my power meters.

The one thing I wanted to avoid was adding any more probes & transitions to increase losses and aggravate the moding situation. I came up with this approach that is very crude but effective.

The basic concept is to have the circular waveguide slip concentrically over the centre pin of the male type N connector, without touching it. I did this by cutting a Type N double Female (bullet) in half and drilling out the centre to hold the outside of the .188 ID tubing. The tube is located so that it is concentric with the centre pin and set in length to just seat against the rear of the connector on the power meter when tightened by hand.

I have no access for equipment to sweep frequency to see the effects of moding so here are my findings at 47GHz.

The Gigatronics reads 10dB low, the HP head is closer to 20dB low and the PMI doesn't respond at all (>20 dB low). The Gigatronics responds at 76GHz as well but I currently have no way to check the calibration.

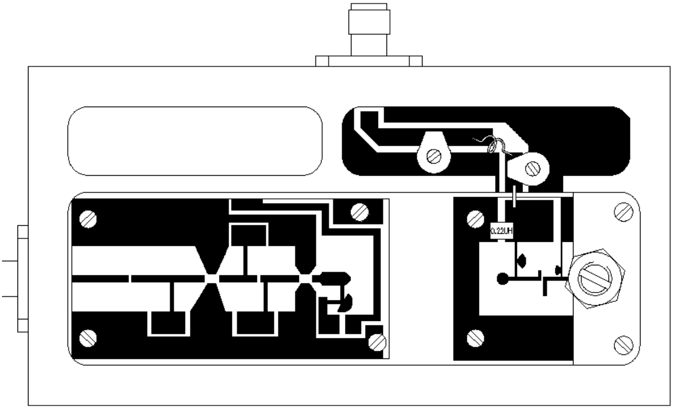
Hopefully this will encourage others to try their power meters at millimetre wave frequencies.

Chapter 7

Transverters and Converters

Improved I.F Matching at 1296MHz for the DB6NT 47GHz Transverter

Paul Widger, G0HNW and Martin Farmer, G7MRF



Following early tests on 47GHz, with the matching network (see figure 1 above) as described in the last issue of the Microwave Newsletter, our "group" in the North West decided that, whilst the circuit did work, it was not very good on receive. This was evident when David,G8VZT/P and Martin, G7MRF/P, were at the Stiperstones in Shropshire during the 24/47GHz contest. Signals from David, G0IVA/P, at Merryton Low were very strong on the transverter that had been modified as described below but on the transverter with the original matching circuit G0IVA's signal could not be heard! Paul, G0HNW, did some bench work on the original matching network and found it would not perform to his liking – so out came the knife and Paul's theory about the pad on the IF being too big at 1296MHz was put to the test.

The pad and the coil seem to be resonant at about 139MHz in their original configuration (figure 2) so G0HNW cut the pad on his PCB to suit the direct attachment of.085 semi-rigid coax cable. The modified pad is shown in figure 3 on the next page.

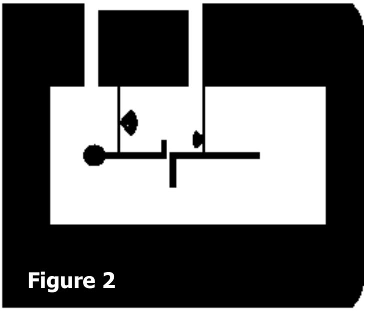


Figure 2

As can be seen in figure 4, the semi-rigid cable passes through the wall of the machined case and is held in position by a small, thin piece of brass shim soldered to the outer casing of the cable and then fastened down with the PCB screw. Paul used a small solder tag to achieve the same result.

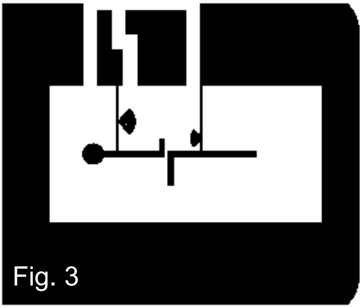


Fig. 3

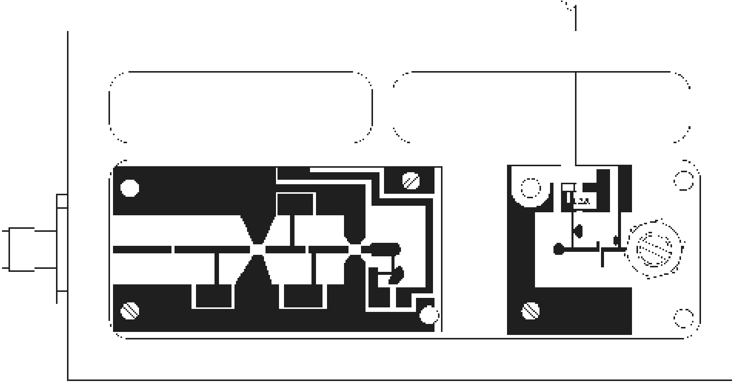
It was thought that the 15 milliwatts of 1296MHz drive required in the DB6NT design was too high a level for the new layout. After some tests by G0HNW, a value of 1.5milliwatts was arrived at.

G7MRF has also modified his transverter in line with this and also ended up by reducing the drive level, down to 4.5miliwatts. This was achieved by G0IVA/P watching the S-meter on his receiver over the 90km path being worked at the time while a variable attenuator was used on the IF transmit line to determine the best level.

The original matching circuit will be returned to for further investigation in the future when time allows.

Following the modification, G8VZT was able to take signals over a 133K path whereas before, over 90Km, nothing could be heard!

Fig. 4



G3WDG040 S-Band Downconverter

Charlie Suckling, G3WDG

Introduction

The G3WDG040 S-Band downconverter is a very low noise figure downconverter from 2.4GHz to 144MHz. It incorporates a HEMT LNA and modamp second RF stage, an image rejection filter, sub harmonic mixer, IF amplifier and local oscillator.

The design is based on two already well-proven designs, the WDG010 2.3GHz transverter and WDG016 local oscillator, the latter being a development of the G4DDK001 design. The HEMT LNA is based closely on the excellent DJ9BV preamp. The HEMT preamp stage is unusual in that the HEMT operated with zero volts on the gate.

It has been found that, with the WDG010 transverter (and WDG025 and 026 preamps), the optimum noise figure for the HEMT at this frequency occurs when the gate voltage is only very slightly negative and that the NF with 0V on the gate is only a few hundredths of a dB higher. Therefore it was decided to dispense with the negative voltage generator used in the other designs to save complexity and reduce cost. The prototype converters have an overall noise figure of about 0.65dB, with a conversion gain of about 38dB. The high gain will allow a reasonable amount of cable loss between the converter and the 144MHz receiver without significantly degrading the system noise figure. In case it is felt that this gain is excessive, it ought to be possible to leave out the modamp IF amplifier, and bridge the input to output with a piece of wire. Be prepared for a possible loss in overall receiver sensitivity if this is done, however, depending on individual circumstances.

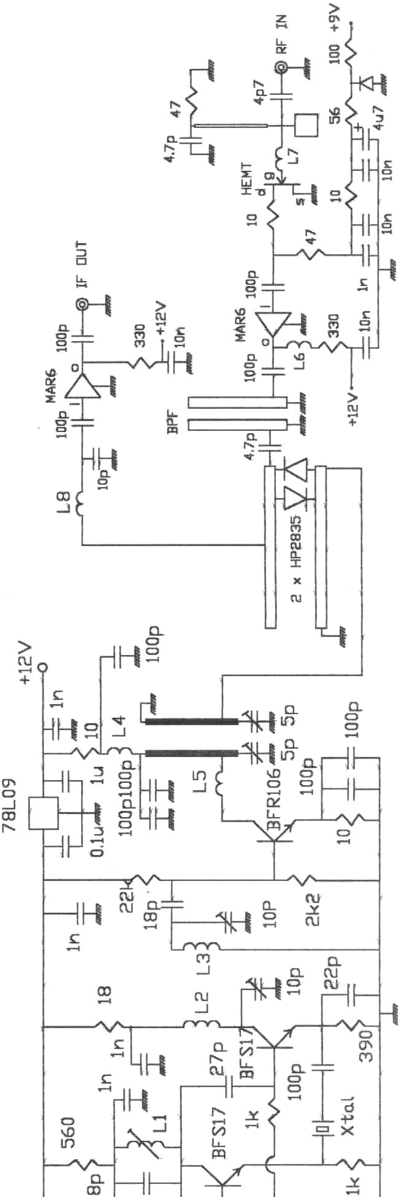
Provision is made on the board for the DC power to be fed up the cable. A bulletin to anyone interested in trying this on request. The extra components to do this will be provided separately.

Construction

The circuit diagram and the parts layout is shown on the following page. Construction follows the general methods used for previous WDG modules . Special points to note are that coils L2 and L3 need to be wound and shaped so that when fitted to the board they are nearly touching, in line, at the same height and with closely spaced turns. It is best to shape the coils and bend the leads before soldering them in place as the tracks may be lifted off the board if too much mechanical force is used later. Also, locate the two 5pF trimmers so that they are touching and position the oscillator coil so that the indent in the top plastic is to the right (i.e. nearest the voltage regulator).

The crystal is mounted through the board and the can is not to be soldered to ground or touching the ground plane. The leads of the oscillator coil and the trimmer capacitors are bent outwards so that the components sit flat on the board.

The choice of RF connectors is left to the constructor. The prototypes used gold plated SMA types soldered to the box. SMA types have the advantage of being not only first-class RF connectors but also do not require much mechanical force to mate. This avoids the possibility of flexing the box excessively when mating/de-mating connections, leading to possible damage to the unit (e.g. cracked chip capacitors). If the unit is to be mounted in a waterproof box later, it may be


78L09
+12V
1n
10
L4
1u
0.1u
100p100p
100p
L5
5p
5p
BFR106
100p
100p
10
22k
18p
10P
2k2
1n
L3
18
L2
10p
22p
1n
1n
27p
BFS17
BFS17
390
560
L1
18p
470
BFS17
1k
100p
Xtal
1n
1k
1k
820
L8
10p
100p
MAR6
100p
IF OUT
330
+12V
10n
BPF
4.7p
2 x HP2835
100p
MAR6
L6
330
10n
100p
47
1n
10n
+12V
10
HEMT
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L7
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10n
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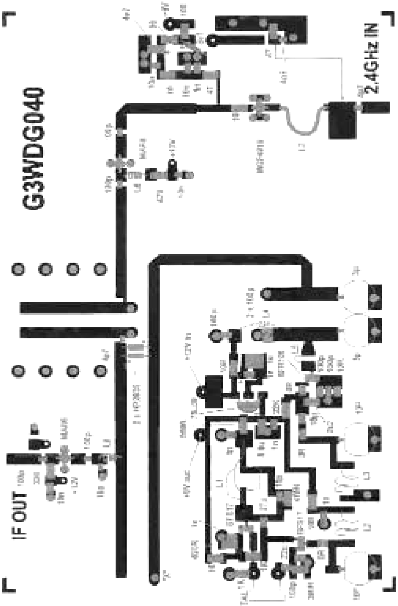
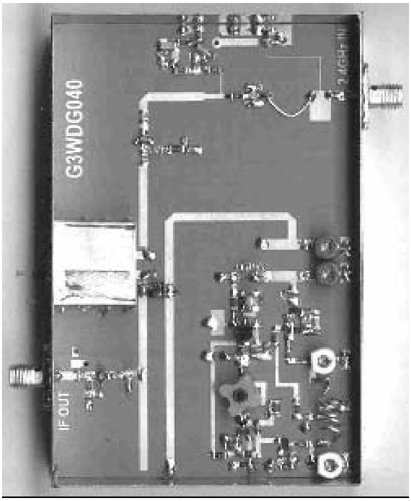
47
4p7
RF IN
56
4u7
100
+9V


G3WDG040
IF OUT
2.4GHz IN


G3WDG040
IF OUT
2.4GHz IN

worth considering connecting coaxial cables directly to the unit. Suitable cables would be thin semi-rigid or ptfe dielectric flexible. These can be connected with the inner passing through a hole in the wall of the box, with the outer folded back at 90 degrees to the cable, strand by strand (to form a disc not twisted together as one) and soldered to the outer of the box. Suitable chassis-mounting connectors fixed to the walls of the waterproof box are then fitted to the other end(s) of the cables.

A photo of the track side showing the components and a component placement diagram are shown alongside the circuit diagram.

Tuning-up

The first step is to tune up the LO chain. Preset the core in the oscillator coil to be level with the top of the former. Take care not to crack the core. Proper trimming tools for these cores are very expensive. I use a ceramic bladed screwdriver which has a blade which just fits across the diagonal of the square hole, and have only broken one core to date. Preset the trimmers to the positions shown in the photo (note the white blobs in the figure correspond with the square indentations in the top of the trimmer where it is adjusted). Apply +12V to the input of the voltage regulator, leaving all other power wiring as yet not connected. The current drawn will probably be in the region of 25-30mA. Adjust the left hand white trimmer (the one connected to L2) slightly to peak the current, followed by the other white trimmer and then the oscillator coil core.

It should be possible, after retuning a few times, to get the current up to near 50mA. We recommend using a ceramic bladed trimming tool to adjust the trimmers to avoid detuning by the trimming tool. If such a tool is not available it may be possible to trim a matchstick or similar insulating material to fit the top of the trimmer. The use of a metal-bladed tool will cause severe detuning. If 50mA current is not achieved, try adjusting the spacing between L2 and L3 (keeping the turn spacing tight) followed by readjusting the 10pF trimmers to peak the current as far as it will go.

Switch off and complete all the underside wiring. Connect the 144MHz receiver to the IF out socket and a good 50 ohm load (matched load, 10dB or greater attenuator or a long length of lossy cable) to the RF input. On switch on, there should be a considerable increase in noise at 144MHz.

Some or all of this will be coming from the IF amplifier. Next, adjust the blue trimmers slightly to peak the noise. This should get the output filter in the LO on tune. Once the LO is running properly there should be a lot more noise than with the LO off tune, and the noise should drop very noticeably if you put your finger on top of the RF filter lines. If a noise figure meter is available, the noise figure of the unit can be optimised by bending L7 up and down (careful not to lift the tracks), or by adding tuning tabs to the wide line in the input, on the opposite end to where L7 is joined.

Tuning adjustments are unlikely to improve the NF by more than .05dB, so don’t be too concerned if you are not able to do this!

This completes the adjustment of the unit. If things don’t work out, don’t despair! Troubleshooting methods might include using an FW radio to check that the 94MHz oscillator is running, loop/diode detectors to look for RF around L2/L3 and possibly the output LO filter. For reference, the crystal runs at 94MHz, L2 and L3 are tuned to 282MHz, and the final filter is tuned to 1128MHz. The final doubling takes place in the mixer. The BFR106 thus acts as a x4 multiplier and produces (in the two prototypes at least) about -1dBm output power at 1128MHz. The mixer needs -3dBm to 0dBm for best operation, although it will probably work well over a wider range of LO levels. It is hoped that the -1dBm level the LO seems to generate will leave enough margin to make this a relatively uncritical design. The most likely cause of poor performance is low LO power. If suitable test equipment is available, the first stage in diagnosing problems is to check out the actual LO level being generated.

The easiest way to do this is to break the track leading from the final LO filter to the mixer by cutting a small gap with a scalpel blade and solder a piece of thin coax (flexible, or 085 semi-rigid) to the microstrip line connected to a power meter. The ground is provided by drilling one or two holes in the ground plane no more than 2mm away from the line, and inserting Vero pins or similar soldered to the ground plane. Later revisions of the PCB have this ground pad in place. After checking the LO level and making adjustments to bring it to the correct level, the coax is removed and the gap repaired with a piece of thin brass or copper foil.

Final Checks

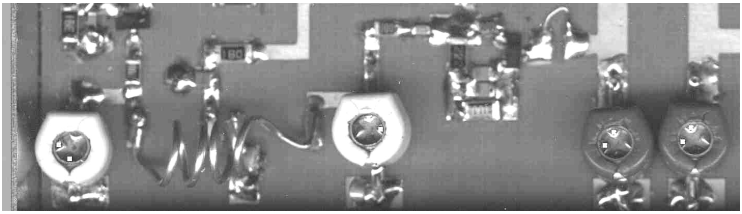
The best way of checking that the downconverter is working is to test it on air. There are two ways of doing this which do not require any specialised test equipment. Both require the use of an antenna.

The first way of checking the unit only needs a low-gain antenna such as a small helix (e.g. G3RUH dish feed). Take the downconverter and receiver outside where you have a good clear view of the sky overhead. Connect the antenna to the downconverter and listen for a change in receiver noise level as the antenna is pointed first towards the ground (the antenna should be kept a few feet above ground level to prevent detuning) and then towards the sky. If all is working correctly, there should be a very noticeable drop in noise when the antenna points skywards (5-dB). If little or no change is heard, there is probably something needing investigation. The second way of checking the receiver is to look for sun noise, but this requires a larger antenna such as 60cm (2ft) dish or larger. With this size of dish, a small increase in noise should be heard when the antenna is pointed at the sun compared to at clear sky. Sun noise may be used to optimise the antenna, for example to find the best position for the feed. The sun noise signal level is more constant than from a satellite, albeit a lot weaker. With dishes around 1.2m (4ft) the sun noise should rise to about the same level as the ground noise.

Coil Details

L1 Toko coil (supplied in kit)
L2, 3 2t 0.65mm wire, 4mm inside diameter, approx 2.5mm above board
L4 6t 0.25mm wire, 2mm inside diameter, approx 2mm above board
L5 hairpin of 0.25mm wire, flat on board, size as per photo – no longer than shown!
L6 4t 0.25mm wire, 2mm inside diameter, approx 2mm above board
L7 15.5mm length 0.5mm silver plated wire, ends bent down approx 0.5mm, approx 0.5mm
above board (adjusted for best NF). Note: wire supplied is LONGER than this and needs to be cut!
L8 as L6

Note: The easiest way of achieving the correct internal diameter for a coil is to wind it on a former (e.g. a drill bit) with a diameter the same as the required internal diameter.

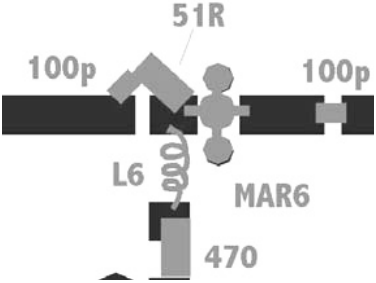


WDG040 Update Bulletin 1 - High Gain "MAR-6"

We have recently come across one example of a WDG040 which had unusually high conversion gain (47dB instead of the usual 39-40dB). This is not a problem in itself, but did lead to lid-induced oscillation problems, requiring the LNA screen described in the Appendix to the construction notes to be fitted. The converter was then stable, but still had the very high gain. [Lid-induced oscillations can manifest themselves as either a large increase or decrease in noise when the lid is fitted, usually accompanied by a very large drop in overall sensitivity].

We have traced the high gain to the MAR-6 device used in the RF amplifier section. After contacting the manufacturer (Mini-Circuits), it seems that the die used in the MAR-6 has been replaced with a different type, which has a lot more gain at 2.4GHz than the previous one. They did this without telling any or all of their customers (thanks Mini- Circuits!). The new MAR-6 have (difficult to read) laser markings on the top of the package, while the previous ones with normal gain have white paint markings. Thus the two types are easily distinguishable.

As noted above, the fitting of the extra screen over the LNA (as well as the one over the filter, which should be fitted as standard) should cure the problem. The screen needs to be soldered to the sidewalls of the box in several places per the photo in the write up. Insufficient soldering points may render the screen ineffective. We have found a simple way of reducing the extra gain back to nominal, should this be required. The "fix" is to add a series 51 ohm 1206 chip resistor to the output of the MAR-6 in the RF amplifier section. By orienting the 100pF and 51R resistors at an angle, the pcb does not need to be modified. The arrangement is shown in the diagram below.



Screens in WDG040 / MAR-6 and MSA06

Someone asked if the "51 ohm resistor" means that the small screen over the filter will not then be required.
The little screen over the filter is still a requirement.
The extra screen over the LNA section noted at the end of the booklet hopefully will then not be needed but be prepared to fit it if the unit is difficult about working with the lid on!

Measurements and Enhancements to the G3WDG 13cm Transverter

Bryan Harber, G8DKK

Background

I have a WDG010, 13cm transverter that has a beta version board. The measured performance of the receiver is excellent: 1.2dB NF (Adastral Park Round Table 1999). On transmit, the beta version board delivers 1mW maximum from the single MAR6 MMIC that follows the mixer + filter combination. Later versions of the board have a second MAR6 to increase the output level to around 3-5mW.

I have the companion WDG035 PA, also the earlier IC based version (now obsolete), that is specified to deliver 1W output from 5mW of drive. The output power of the transmit converter/amplifier combination measured with my professional microwave power meter is 500mW. This indicates the amplifier gain of 27dB is about 4dB greater than predicted. During the process of measuring the 1W output, a disturbing feature was observed: with no 144MHz drive applied the power meter was showing an output power of –7dBm, 34dB below the wanted output power. The 144MHz source had been disconnected and the transverter manually set into the transmit mode to confirm the problem.

Further Measurements

The transverter was taken into my place of work and the transmit output observed on a microwave spectrum analyser. The source of the unwanted output was the 2176MHz LO component that was indeed at –7dBm at the PA output.

The output spectrum is shown in **Figure 1**.

Removing the PA and measuring directly at the transmit output of the transverter module showed the 2176MHz signal was now at –34dBm as expected. A further unwanted component was also observed at 1088MHz but this is more than 50dB below the wanted output and some 20dB below the level of the 2176MHz signal.

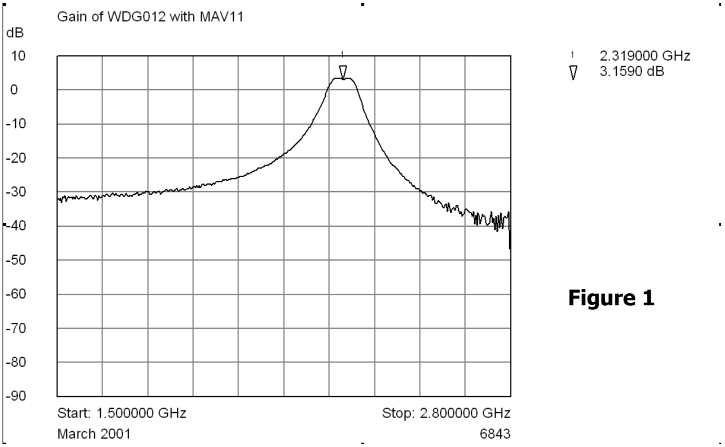


Figure 1

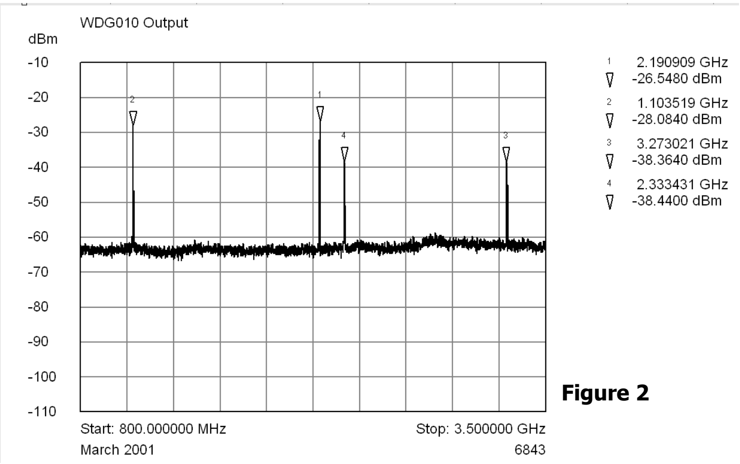


Figure 2

These results were entirely compatible with the design that uses just a single filter, consisting of 2 coupled lines, following the mixer.

Enhancement Solution

A few dB of extra gain is required to drive the PA to full power and an additional 20dB of filtering of the LO signal would reduce the LO leak signal.

Still in the Microwave Components Committee development list is the WDG012 filter/amplifier module. This module is configured with a filter identical to that in the transverter followed by a MMIC amplifier, an MAV11, contained in the smallest (37mm x 37mm) tin-plate box. I obtained one of these modules and completed it in 2 hours.

The measured response is shown in **Figure 2** above. A gain of 3.15dB was obtained with an input power of approximately 1mW that represents an output power of +3.5dBm or 2.25mW. As can be seen in the plot the marker at 2176MHz shows that the filter provides a rejection of some 20dB compared with the level at 2320MHz. The filter also gives around 50dB rejection at the LO sub-harmonic frequency of 1088MHz. Connecting the WDG012 module to the transmit output of the transverter provides a much improved spectrum, in fact I was now unable to measure the 1088MHz signal and the 2176MHz signal has dropped to a very satisfactory 35dB below the wanted signal level.

In the spectrum shown in **Figure 3**, the spectral line at 2.320GHz (marker value 2.333431GHz) is shown with the applied drive level 30dB lower than that normally applied by the 144MHz transceiver.

Further Enhancement

All looked good but I was surprised that the MAV11 provided only just more than 3dB of gain until I studied the manufacturer's data for this device. It is rated to only 1GHz and although I may not have an ideal grounding of the device it explains why the gain is lower than expected, although adequate in this application. The output power and output –1dB gain compression point are also satisfactory at 2.25mW and +12dBm respectively.

I had recently come by some of the newer ERA series of MMIC amplifiers and decided to try the ERA4. This

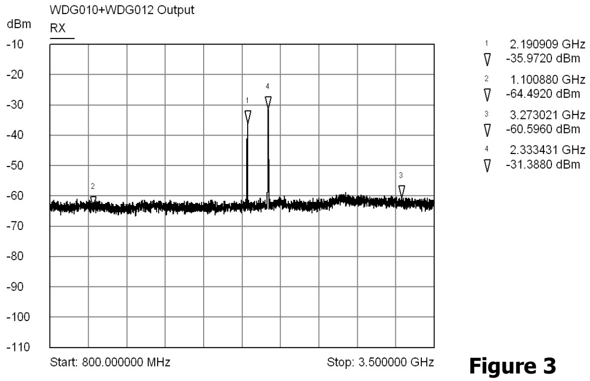


Figure 3

device is rated up to 4GHz with a –1dB compression point of +15dBm. I exchanged the MAV11 for an ERA4 and re-measured the amplifier with the following results:

Gain is now 9dB at 0dBm (1mW) of drive giving an output power of 8mW and the –1dB output gain compression point measured as +15dBm (30mW) with an input power of +7dBm (5mW). All of the filter characteristics remain unchanged and no sign of spurious oscillation was observed on the spectrum analyser. The measured response of the WDG012 with ERA4 is shown overleaf in **Figure 4**.

Final Results

The completed modules are now assembled in a waterproof ABS box that is mounted just below my 13cm Yagi antenna. The completed transverter consists of: WDG010 transverter module, WDG012 amplifier/filter fitted with ERA4, WDG035 PA (early IC based version, now obsolete). The final output power of the combination of transmit converter + amplifiers measured with my microwave power meter is 1.1W.

Conclusion

The WDG012 module provides worthwhile additional filtering to the WDG010, 13cm transverter in that

it reduces the output levels of unwanted products that will otherwise be amplified by the following broadband amplifier stage (s). Fitting the ERA4 in place of the MAV11 seems to give a worthwhile, although not essential, improvement in the performance of the WDG012 module.

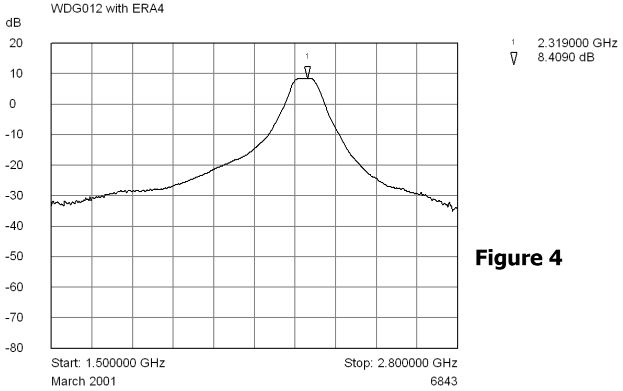


Figure 4

Experiences with the DB6NT 5.7GHz Transverter

Paul Marsh, M0EYT

After having a few years of operation during the 10GHz cumulative contests and being asked, repeatedly, if I was QRV on 5.7GHz, I decided that it would be a good idea to get on with building up a transverter for this band.

I had tried in vain to tidy out the junk in my garage and decided that, before skipping the stuff, it should go on EBay. Amazingly, the 'rubbish' that I'd cleared sold at good prices and amounted to enough to purchase a 5.7GHz transverter kit. Our local radio club (**www.frars.org.uk**) and another local radio amateur were also interested in getting some Kuhne Electronics (DB6NT) hardware, so we put in a combined order for 4 transverters covering 2.3GHz and 5.7GHz. After placing the order with Kuhne, it was only a few days before the large box arrived, very well packed and complete with a few of the new catalogues. I started the 5.7GHz transverter construction at once. The instruction / build manual is very straightforward and is well worth a couple of passes before gluing all the hardware together.

The tin plate box needed to be drilled and tapped to accommodate the 3 SMA connectors and 4 DC feed through capacitors. In my transverter I added another SMA socket for external LO input as I was intent on using an external G8ACE type local oscillator. The mechanics took about an hour to complete, including soldering the two halves of the tin plate box together and installing the RO4003 Duroid PCB. If you've not seen a DB6NT transverter kit before, they are nicely presented with all components being supplied in compartmentalised plastic boxes with all values printed on the lid of the boxes. Two boxes are supplied for the 5.7GHz transverter, one containing all the Rs and Cs, the other containing semiconductors, sockets and other mounting hardware. The transverter was built up as per the directions in the manual, starting with the two resonator cavities, helical filters and the discrete components. The power supply semiconductors were installed next, and a DC check performed to make sure the voltages looked right, which they all did.

The MMICs and FETs were added without problem – well almost with out any problem! Two transistors in the multiplier chain caught me out with the orientation of the writing on top of them being 180 degrees out compared to the component placement diagram but it was simple to rectify. I should have read the note on the component placement diagram which says "Don't use the orientation of the writing as a guide!!!"

I found it useful to use a good temperature controlled soldering iron for all the steps. The iron and work area was prepared to minimise the possibility of static build up, which could damage the devices.

The next step was the tune up as per the manual. The local oscillator is fairly straight forward, with several multiplication steps from the temperature-stabilised crystal of 117MHz up to the local oscillator of 5.616GHz. The tune up steps recommended measuring voltages at various points along the chain, and peaking for maximum / minimum. This was done and the results checked on a spectrum analyser only to find that no further improvement could be obtained. The remaining cavity filters were tuned as per the instruction – connect a 144MHz receiver and listen for a peak in the noise. The transmit side was pretty similar; apply 144MHz RF and peak for maximum power output, using the built in power monitoring facility. The transverter was checked on air and the local GB3SCC beacon could be heard on about the right frequency.

The next stage was to integrate the transverter into a complete system that I could take out on the local hills. For my portable systems, most of the transverters are housed in lightweight plastic boxes, with an aluminium chassis inside. The boxes used are made in the UK by Gewis (**www.gewiss.com**) and are of type GW44208 IP56.

Before getting the transverter kit, I had purchased a 5.7GHz power amplifier strip from another UK amateur – it was a TWT replacement and as such had a pretty high gain set of amplifier stages. This PA (see Fig. 1 below) was to be incorporated into the final transverter design, so the first two stages were removed, giving a PA that needed about 100mW in for 5W out – I built up a simple protected power supply circuit as a module and put that to one side whilst I concentrated on the rest of the transverter integration. As with any 'high power' microwave transverter, sequencing is recommended to avoid having to replace front-end devices repeatedly. I chose to use the VK5EME EME66 sequencer which has 4 sequenced outputs, and given the current £ to AU$ exchange rate, they are very cheap to import.

The sequencer has an output for a receive preamp which gives me the flexibility to add one in the future. The other outputs drive the NMS1212 dc-dc converter for the 24V relay, the antenna change over relay and finally the supply volts to the PA – this was via another small relay to avoid the voltage drop when switched through the sequencers output transistor. I used a latching 4 port transfer relay in my 5.7GHz transverter – it's an R566433 type, brand new from EBay for a "fiver" – the specification seems pretty good and the isolation is sufficient between the ports.

The other addition was a DF9LN local oscillator module that I bought from the 'flea market' at a Crawley microwave round table meeting – this really has improved the stability of the transverter. The picture below shows the completed transverter with the temporary (!) wiring that was installed just to test it.

I had my first 5.7GHz QSO only a couple of days after completing the transverter; I worked Dave G0RRJ via rain scatter on the 2/10/2004. So far, the system has been used in a few cumulative contests and some of the other

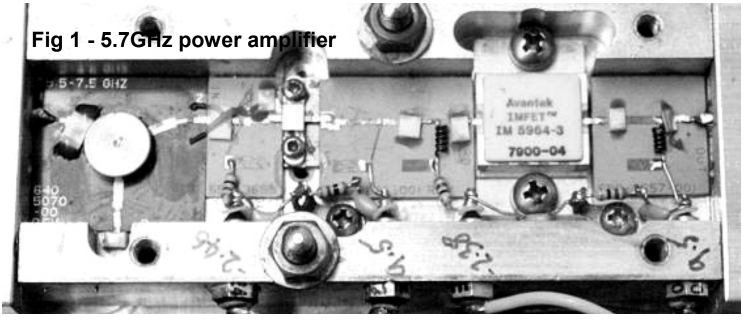


Fig 1 - 5.7GHz power amplifier

VHF/UHF/SHF contests that either FRARS or I participate in. I get the feeling that 5.7GHz seems to go better over paths that you might struggle on when working 10GHz. I've had a few QSOs from home with just a feed-horn propped out of the window, successfully working G0RRJ and G4LDR via some scattering mechanism – tree / rain / other at a guess. Certainly rain scatter effects seem to be more pronounced at 5.7GHz that 10GHz – all we need now are a few more 5.7GHz beacons in the UK with which to test.

Overall, I can thoroughly recommend the DB6NT transverter kits, they are easy to build and guarantee very good results on the microwave bands. Have a look at **www.db6nt.de** for more information on the 5.7GHz transverter – you can buy either a ready built one that has been professionally aligned, or save a bit of money and buy the kit, and have a few enjoyable hours building it up and testing it out on the air.

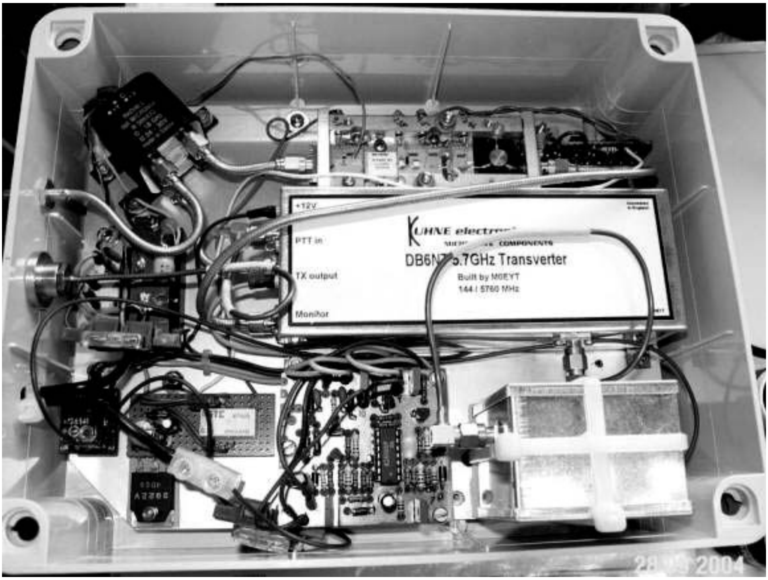


Fig 2 – Completed 5.7GHz transverter

Six Centimetres in two Days!

by Peter Day, G3PHO

I don't know what it was that encouraged me to email Michael Kuhne, DB6NT for one of his 5.7GHz transverter kits but I'm very glad I did! The kit arrived within a few days of ordering and it literally took only two days to build into its tinplate box! A further day was spent making a suitable enclosure for the complete transverter (i.e. including c/o relay, metering and IC202 interface).

On opening the parcel from Kuhne Electronic, I was immediately impressed by the high standard of presentation. There were two plastic, compartmentalised cases containing the components. Each compartment was labelled with the component value.

The construction manual is in English and the circuit, pcb layout and other diagrams, including photographs are beautifully presented. Most of the components are very small SMD types and are a very good test of one's eyesight! The transverter is of single board construction on RO4003 substrate. The DB6NT catalogue guarantees an RF output of more than 200mW. I found it easily did that and much more. A receiver noise of 1dB is claimed but I have not had the opportunity to check this yet. I will have to wait for the next Martlesham (Adastral Park) Microwave Round Table for that! The pc board is of excellent quality, gold flashed on the ground plane side and with plated-through ground connections from the circuit side. It fits into a tinplate enclosure 33 x 55 x 150mm. SMA connectors are used for IF input and the separate antenna connections to RX and TX. Feedthroughs are used for the 12V DC input line, PTT, power monitor and the +12V output for external relay control. Everything is provided except for the external c/o relay at 5.7GHz. As received from DB6NT, the transverter is configured for a 144MHz IF but, with some slight component changes, it can be used with a 432MHz IF. State which you want if you order the kit.

The circuit has an onboard Butler type local oscillator which I found to be very stable once the warm-up period had expired. The crystal uses a 40 degree PTC heater to make this period as short as possible. An external, high accuracy 117MHz source such as the DF9LN LO, can also be fed in if need be. The 117MHz is tripled in a BFR92A and then doubled in a BFP196 to 702MHz. A second BFP196 doubles again to 1404MHz and an MGF1902 quadrupler stage takes it to 5616MHz. Helical filters are used at 351MHz, 702MHz and 1404MHz to provide very good rejection of spurii. A microstrip edge coupled filter selects the 5616 signal to drive an ERA 2 MMIC to 5mW out put. The mixer is a single balanced diode type (BAT15-99 double array). Pin diode switching takes care of the TX/RX side of things apart from the antenna changeover, where a coax relay is required. The Receiver section uses a two stage HEMT amplifier

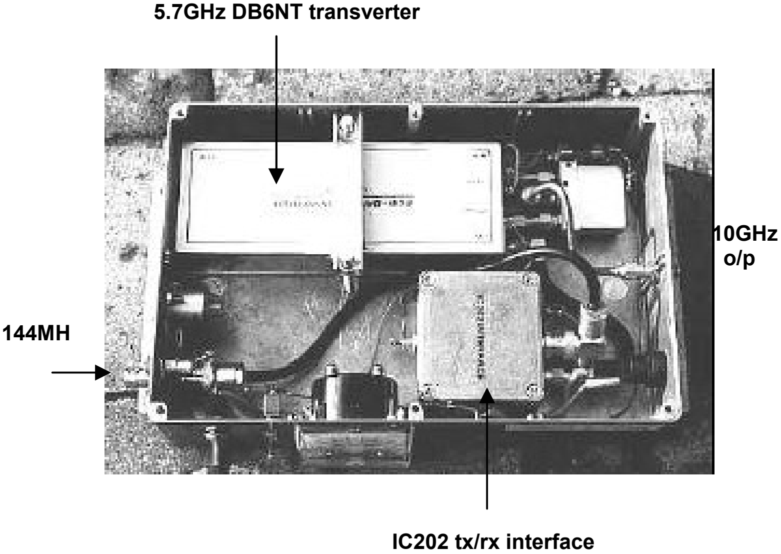
(2 x NE32584C) followed by an ERA3. A Wilkinson divider couples this to the mixer. The Transmitter has two stages of MGF1902s after the Wilkinson divider to drive an MGF1601 to over 200mW out. A simple power monitor or using a BAT15-03W diode gives a good indication of RF out. Cavity filters are used to provide the necessary LO and image rejection for the RX and TX. It had been some time since I had tackled a job of this nature, though in recent years I have been experimenting with SMT devices and smaller microwave modules such as the DB6NT 24GHz mixer and various WDG kits. Michael states in the manual that this kit is NOT a beginner's project and he is quite correct. However, anyone who has built up WDG and DDK modules in the past should have few, if any, problems. Essential equipment, in my view, is a temperature-controlled soldering iron (a Vann Draper SL20 in my case) with a fine tip, binocular headset magnifier (RS part no. 606-589 in my case) or large magnifying glass, a good pair of fine tip tweezers and good lighting! It goes without saying that modern surface mount components are very small and easily lost if allowed to snap out of your tweezers! Of course, you must also take the usual ESD precautions when handling the static sensitive devices. The actual construction presented no problems. Before the active devices were installed the pcb was checked for soldering errors. The voltage regulators were then installed and checked for the correct output. Then the bipolar devices were fitted and voltages checked. The LO was checked against my frequency counter and adjusted to 117MHz. Finally, the GaAs-FETs were installed and bias levels checked. After that the tune up was very straight forward since Michael provides clear instructions with relevant voltage levels for the various test points. The filters tuned up beautifully. No one was more surprised than I was when the marker from my Adret 5401 synthesiser/diode multiplier arrangement was heard very loud and clear at T9, bang on 5760.100MHz! I use IC202s for my microwave IFs. Both have been modified to give 120mW out

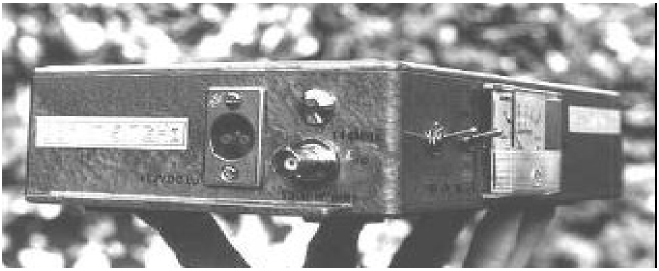
instead of the more usual 3 watts. Since the DB6NT kit comes with an IF input attenuator for a 3W input, I had to modify this to allow the lower input value to drive the unit. This meant reducing the 470 ohm resistor in the PI-attenuator to 100 ohms.

The fitted variable 100 ohm pot could then be adjusted for correct IF drive to the mixer.

The DB6NT transverter, as it stands, is configured to use an FT290 prime mover where +9 volts or so is present on the coax during transmit. This voltage switches a keying transistor into TX mode. I decided to leave the DB6NT intact and to make a simple two transistor keying interface for the IC202 so that, on pressing the microphone switch, the interface would ground the DB6NT PTT terminal, thus switching the module to TX. The IC202 has around +4.5 volts present on the coax inner on RX mode rather than the TX mode of the FT290. The interface was built into a small die cast box.

The tinplate box housing the transverter, the IC202 interface module and the microwave coaxial change-over relay were then installed into an Eddystone die cast box (220mm x 145mm x 55mm) which had been pre - drilled and sprayed with silver hammer finish paint. Switches, input and output connectors plus a meter to monitor both the 12 volt DC input and the TX RF output where also fitted. The photographs on the following page show the details. The 12 volt change-over relay came from a Teletra 23GHz TX/ RX unit.





The DB6NT Module makes for a very compact transverter … almost a handheld!

Chapter 8

IF Rigs

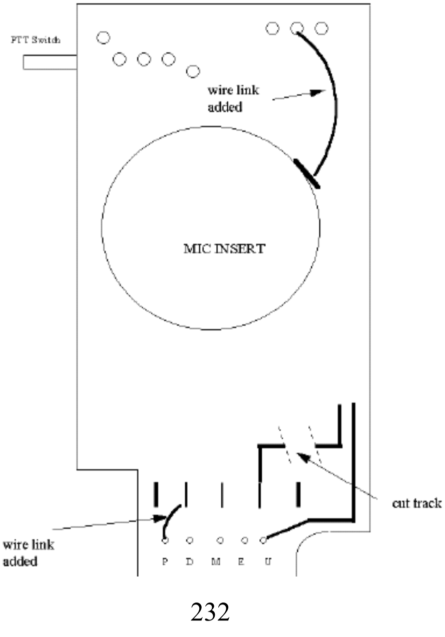
Yaesu FT290 Microphone Mod

Chris Towns, G8BKE

Many of us use the FT290 as a prime mover on the microwave bands. Quite often, during initial line up testing, one is asked to transmit a continuous carrier for perhaps some minutes while the far end searches for your signal. This usually entails holding the PTT or key down for this period. A good strong elastic band is useful for this purpose but that can either get forgotten, or worse...snap!

The Yaesu YM-47 microphone has a slide switch on its rear which allows one to "lock" the up/down buttons. I, like many others no doubt, never use this function and thus decided to use the slide switch for a PTT function. Unplug the microphone from the rig and take out the two screws which hold the microphone casing together. Carefully separate the two parts of the microphone casing and remove the pcb with the microphone insert attached from the casing. Don't loose the up/ down buttons or the PTT lever in the process but note that the up/ down switches themselves come away with the pcb. With the microphone insert facing you, locate the pcb track shown in the drawing. (Not all the pcb tracks are shown for clarity.) Carefully cut the pcb track and add the 2 wire links shown, using insulated wire. Re-assemble the microphone and there you have it. The up/down function is retained but the slide switch will now work in parallel with the PTT switch. Slide the switch to the "on" position and you can send continuous carrier for as long as the other station needs it while you can get on with your ham sandwich!

The usual disclaimers regarding modification of equipment apply.



Using the FT817 in a Microwave System

More gleanings from the internet reflectors

Tom Williams' (WA1MBA) internet microwave reflector is a most useful source of information.

There was much discussion when it was launched, of the value of the FT817 QRP portable Transceiver to the microwave operator. With our thanks to the various operators mentioned below, we publish a summary of this discussion in the hope that it might help those who are looking for a substitute for their IC202 microwave IF "prime mover" or even very portable talkback equipment.

From: Ed Cole, AL7EB,
[al7eb@ptialaska.net]

I have used the FT-817 with my DEM 10GHz transverter and 1w Qualcomm amp. I mounted everything in a U-shaped aluminium chassis which fits into a sturdy carrying case (Pelican-1520) making it handy for taking on-board an airplane when I travel. The case has room for my key, microphone, a 12v gel-cell battery, 1.5 amp wall charger, a 17dBi horn and assorted cables.

Since most of the equipment is pre-mounted on the chassis, set up takes only about 3 minutes using the horn. I use a DEM sequencer to control the transverter and amp T/R as well as controlling microphone PTT. I run the microphone PTT line through one of the sequencer relays to prevent transmitting before the T/R sequence is complete. The microphone keys the control line of the sequencer. To break out the microphone PTT line, I use a couple of RJ-45 wall jack modules and a RJ-45 jumper to the radio microphone jack. For CW I chose manually switched T/R rather than break-in operation to keep it fool-proof and simple. The Qualcomm PSU and sequencer are housed in a plastic box. I use the sma-relay supplied by DEM and have the 3w transceiver option for connecting the FT-817 antenna to the transverter. I adjusted my transverter drive level to use the FT-817 1w power level as this permits me to operate the FT-817 using the battery at lower current draw.

Lower power is easy by switching the FT-817 to the 1/2w power level. I can take along a mobile mag-mount whip or use my Arrow (2m 3 elem/70cm 7-ele) Yagi for liaison. The Arrow breaks down and fits into a 2-inch diameter by 25-inch long plastic drain tube, using plastic end-caps. It looks just like those carry-on fishing-pole tubes.

The equipment sits on the car seat very nicely for mobile operation (I use a cigarette-lighter power cord).

The FT817 will make a very nice IF unit for portable microwave operation, as long as you have adequate battery capacity. The only negative I see is the <relatively> high current draw in receive (~300 mA)

From: Peter Freeman, VK3KAI
[peter.freeman@sci.monash.edu.au]

I have been using an FT817 as an IF box since October and it's an excellent box for the job! The biggest problem is remembering what the IF frequency should be - I have some transverters set up for 144MHz IF, some on 145MHz IF and 2403 set up for 147MHz IF ! During a Field Day, it is easy to NOT change to the appropriate IF.

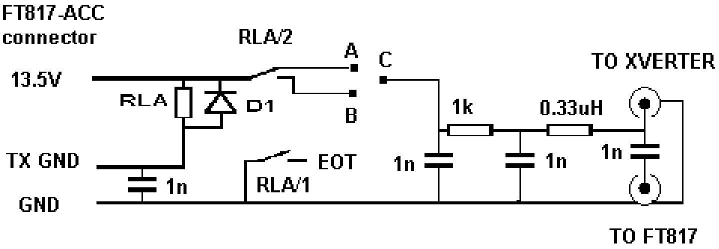
I have built a Bias Tee box that plugs into the RF connector on the rear of the unit and brings in the Tx GND connection from the ACC connector. The RF coax cable then goes to

the IF interface board and controls the sequencing in the transverter.
Another useful item is the keyer … hook up a switch to the keyer plug on the rear of the radio, switch to CW mode, set to Break-In operation and you have a beacon keyer running at 50% duty cycle!

From:WW2R (G4FRE) David Robinson [robinda@nortelnetworks.com]
I have been using one as a microwave IF on 10/24/47GHz since they dropped $100 off the price last year (August). I have been very pleased with it as an IF, especially with the extended frequency coverage, its quite fun as an hf rig as well! I built a bias box for the rear SO239 connector that plugs into the accessory socket and makes the 817 look like an FT290/IC202. I learnt the hard way that you do not have a second chance if you short out the 13.8v output on the acc socket ...it is fed through a 10 ohm 1/16w 0604 chip resistor that blows.
From experience it is a devil to replace.
I got over the small frequency etc display by installing FT 817 commander by HB9DRV that remotes everything onto a laptop.

Interface circuits for the FT817

Dave Robinson WW2R



Interfacing to a transverter:

Refer to the diagram above ...

For IC202: connect points A and C

For IC402/FT290: connect points B and C

This unit simulates either an IC202 (9V on RX) or IC402/FT290 (9V on TX). For RLA I used a very miniature DPDT relay (the antenna changeover relay from a Pye PF2UH!), mounting the components in a small die cast box with a miniature 6 pin DIN socket for the FT817 connections. I then have a lead from the accessory socket of the FT817 to the interface box. The advantage of this is that I can use the same box with a miniature 6 pin din to 13 pin din lead and use it to interface with an IC706mkIIG. The input and output connectors are UHF. The box also produces a ground on transmit signal on a 3.5mm socket to control other circuits.

When the unit is completed make doubly sure there are no shorts on the 13.8V pin, as measured at the miniature 8 pin din aux plug that goes into the FT817, unless you have 40/40 vision and would like to learn how to remove the main FT817 PCB to replace an 0604 chip resistor! If you are worried about this possibility, provide the 13.5V from a supply external to the 817.

Computer Interfacing:

Refer to the diagram on the next page

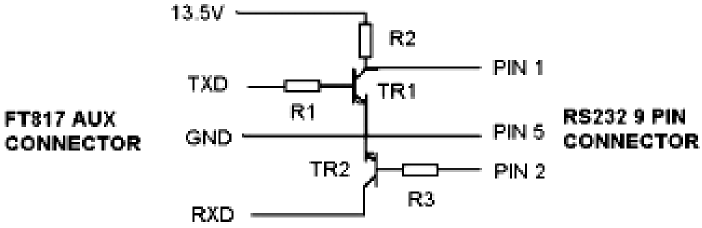
Loading the memories of the FT817 before a contest was a pain, until I discovered the FT817 commander software by Simon Brown HB9DRV (ex GD4ELI!) at:
http://www.kns.ch/sysgem/hb9drv/
All I needed was a cheaper RS232 interface than the Yaesu CT62. I came across a circuit of Russian origin that only required 2 NPN transistors and 3 resistors arranged as on the next page.

The transistor types aren't critical, I just had some 2N3904 around. R1=R3 =39k, R2=3k9. I easily built mine into the 9-pin RS232 connector using 1/8W resistors with a lead connecting to the FT817 auxiliary connector.

Other Notes:

Be aware that, even with the FT817 turned off, there is still 13.5V on the aux connector 13.5V pin.

Don't leave accessories plugged into the aux socket when running off batteries or they will flatten.



For those worried about conserving every last mA of battery drain, see the KA7OEI web site,
http://www.ussc.com/~turner ft817pg.shtml
where the matter is thoroughly investigated, along with lots of other related topics.

My radio bought in the USA was supplied with a ferrite choke but nothing to say why (my addendum sheet was missing). I soon discovered why whilst using it for talkback with W5LUA on 432.1MHz SSB. RF gets into it and it is so distorted as to be unreadable. It doesn't matter if you are using the helical antenna on the front panel or back panel, same effect. I couldn't repeat the effect on any other band/ mode, even the HF bands with the ATX whip. Putting the choke on the external power lead close to the power connector cured it on the Yaesu supplied DC power cord.

Interestingly when powering the radio off an external G3TUX power adaptor (http://www.g3tux.co. uk/) the same ferrite choke doesn't cure it. Having got tired of having to remove the radio from its leather case every time I wanted to plug in his adaptor (because it has a right angled plug which wont fit easily through the hole) I chopped it off and fitted a straight plug with a 100pF chip capacitor soldered across it and that fixed the problem.

Versatile Transverter Interface for the FT817

Peter Day, G3PHO

The FT817 is a deservedly popular transceiver, used by thousands of QRP enthusiasts around the world and immediately recognised by the amateur Microwaver as a modern replacement for the ageing IC202 or FT290 "prime movers" of yesteryear.

I've used a couple of IC202 transceivers (one an S model) for over fifteen years and have been very pleased with their performance as a microwave IF, except for one feature … the poor frequency display of ten kHz per dial division. In the past few years, my microwave transverters have been improved so much, with features such as hi-stability OCXOs, that their frequency stability and resetting accuracy now exceeds that of much of my old VHF gear. Hence I thought it was about time I treated all the transverters to a more modern IF transceiver.

In late March 2004 I purchased a new FT817 and was immediately impressed at how Yaesu could pack so much into so little!

My 3.4, 5.7, 10 and 24GHz transverters are designed around DB6NT transverters, constructed from Michael's excellent kits and driven by 120 milliwatts of 144MHz from an IC202S. My 23cm old DEM transverter is driven at 10 watts input by an aged TS700A, while my recently acquired 47GHz gear, bought from a "retired millimetre man", requires an FT290 at 3 watts. Apart from the 23cm transverter which uses a separate PTT line, all the others require a switching voltage on the coax connecting the IF transceiver to the microwave transverter. The 47GHz transverter send/ receive is controlled by a positive voltage on TX while the others are switched by a positive voltage on RX. A look through my (almost complete) collection of Microwave Update Proceedings found a couple of relevant articles in the 2002 edition one by Paul Wade, W1GHZ and the other by UK expatriate Dave Robinson, WW2R. Dave's article provided me with the solution to the problem of switching the transverters no matter what the configuration (IC202S or FT290).

The final outcome was the unit shown in the photographs. A versatile and switchable interface between the FT817 and the microwave transverters. This little unit allows the FT817 to directly replace either IC202 or FT290 type transceivers with no modifications at all required to the microwave transceivers. A facility to adjust the RF input to the transverters is also included, as well as a near foolproof protection circuit for those who have been known to get the 12V DC input leads the wrong way round … a potentially disastrous event when out / portable!

Apart from a little mechanical ingenuity, most of this article relies entirely on other people's work. My thanks go to Dave, WW2R, for the basic circuitry.

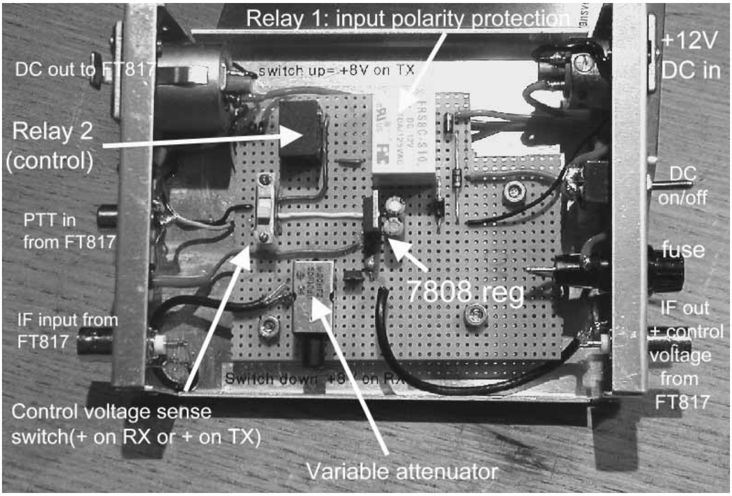
The Design:
The interface module is connected to the +12 volt DC input supply (e.g. a battery if operating portable).

If the supply is connected reversed polarity then Relay 1 does not go over and you will have to reverse the supply leads for correct operation. The correct state is shown by the red LED. Over the years I have found this relay protection just about "idiot proof" and I now have no fear of damaging my precious microwave gear through lack of care when hooking up supplies. The few extra milliamps drawn by the relay is a small price to pay for piece of mind …. and please don't tell me that you have **never** connected a supply the wrong way round!

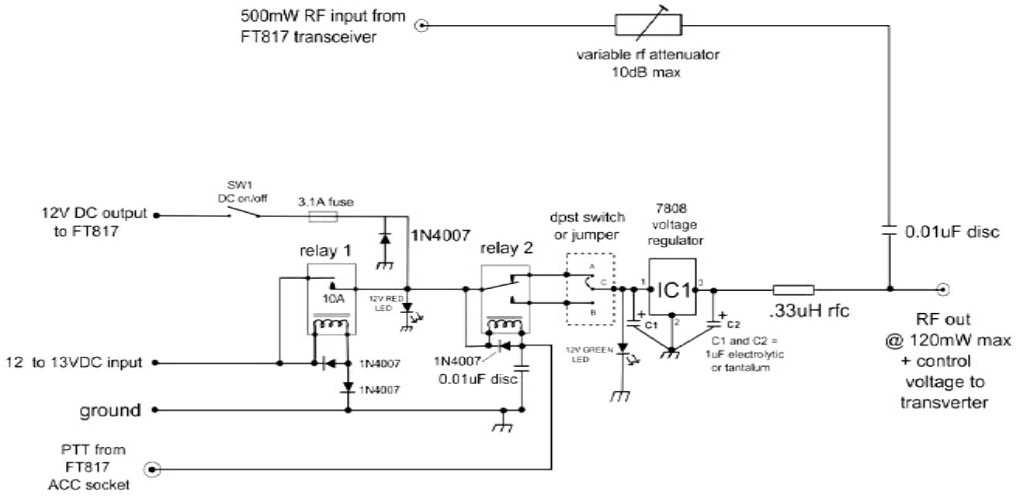
The FT817 power lead is connected to the fused 12V output socket on the interface, while the FT817's BNC VHF antenna connector is connected via a short length of coax to the socket marked '500mW RF input'. The FT817 power output is reduced to 0.5W by means of the transceiver's menu. A coax cable of any convenient length then connects the interface to the microwave transverter. My own system uses a 15 foot length of 50 ohm double screened lead here, the variable attenuator in the interface being adjusted to give 120mW at the far end of this cable.

To obtain full send/receive control of the transverter by the FT817 PTT line, an 8 pin mini DIN plug is wired up to take a short, thin coaxial lead from the transceiver's ACC socket (pin 2) to the PTT phono socket on the interface. When the microphone PTT button is pressed or the Morse key activated, pin 2 ('TX ground') of the ACC socket of the FT817 is grounded to pin 3 and in turn completes the ground connection of the coil of Relay 2 in the interface. This applies 12V to the 7808 regulator IC. Depending on the configuration of the slider switch in the interface, a +8VDC control voltage will be applied to the IF output line on either RX or TX.

Construction:
There's nothing special about this module. One off units like this don't need a purpose made PCB unless you like your gear to look fancy! The enclosure is a small aluminium box I had in the



FT817 /Microwave Transverter Interface



Peter Day,G3PHO, April 2004

“come handy drawer” and I used some perforated board that has strips of copper foil on one side (called “Veroboard” in the UK) for the circuitry. The board should be populated with the various components and tested before it is put into the small aluminium enclosure, as shown in the photographs.

The box is drilled at both front and rear aprons to take the various connectors and switches, making sure that all I/O leads to the FT817 go to one apron and the outputs to the microwave transceiver to the other.

The finished product:
The complete interface is a little smaller than the FT817. If a larger enclosure were used, extra DC and IF output sockets could be fitted, enabling use with several transverters at the same time.

After completing this project I realised it would have been more useful to have the control voltage sense switch on the front apron of the enclosure rather than inside the box!

Components:
The component values are marked on the schematic but a few extra points are added below ....

The variable attenuator:
I already had a few of these nice little pc mounting variables in my junk box. Although they do not provide a constant input and output impedance (and therefore not a good VSWR for the FT817), they are very convenient. A fixed resistive attenuator (T or Pi type) can easily be substituted and calculated from the values of input and output required. I decided not to switch out the attenuator when receiving as my transverters have more than adequate RX gain in reserve to overcome the 6dB or so of attenuation in the IF line. This is probably the case for most systems.

Relay 1:
This is for polarity reversal protection. Any 12V DC type will do here so long as the contacts can handle a few amps. I used one rated at 10A. To reduce overall load on portable batteries choose one with low coil current.

Relay 2:
This is the main control relay. I used a miniature 12VDC type with 1 amp contacts.

With both relays, don’t forget to put 1N4001, or similar diodes, across the coils to prevent “back EMF” effects.

RF choke:
A small axial lead 0.33uH type

8 Pin mini-DIN connector (male):
This is needed for connecting the interface to the ACC socket on the rear of the FT817.

Make sure that the inner of the thin coax lead goes to pin 2 and the shield to pin 3 of the male connector. Use the pin numbers on the connector itself to absolutely sure you have the correct pins. Wrap some insulation over nearby pins, one of which, pin 1, carries +13.8V from the T817. Short this to ground at your peril! Refer to the FT817 manual diagram. On the ACC socket pin 2 is named “TX ground” and pin 3 ground”.

Do not use the PTT pin on the DATA socket.

Control Voltage Sense Switch:
A small slider switched was mounted

on three pins that are soldered through the perforated board.
Pin 1 of the 7808 regulator is connected to the slider. In the schematic, pins A and C are connected by the switch to simulate IC202 control switching while pins C and B provide FT290 type switching. This switch could be mounted on the front panel for ease of use.

Internal IF coax lead:

I used a short length of RG174 here, taking care to minimise the exposed sections of inner conductor at the attenuator and the BNC connectors.

Summary:

The interface worked first time and has been used with excellent results in the May and June 5.7/10GHz cumulatives and millimetre band contests. One added advantage of using the FT817 is that if ‘KYR’ is selected by pressing key F on the front panel, the C function switch below the main dial puts the FT817 into beacon mode, sending a steady stream of dots or dashes according to your tastes (provided a Morse key is plugged in).
After over a decade of using IC202s to make hundreds of microwave contacts you get quite used to using antiquated gear. Only now do I feel I’m in real control of frequency setting. I got quite a “buzz” from a report by G3LRP that I was only 100Hz out on 24GHz when I called him on 24048.100MHz. I can live with that considering I do not yet have GPS locked oscillators!

References:

1. Bias Tee for Remote Transverter Control - Paul Wade, W1GHZ: Microwave Update Proceedings 2002

2. Interface circuits for the FT817 - Dave Robinson, WW2R (G4FRE): Microwave Update Proceedings2002 and also RSGB Microwave Newsletter April 2002

Power Reduction for the FT817

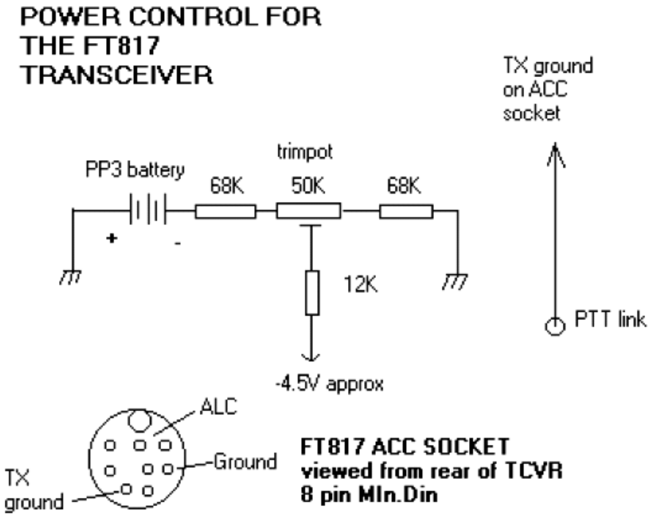
Martyn Vincent, G3UKV

Editor's note:

A number of modern transceivers while allowing you to turn the power down to 5 Watts or less, often have an initial "spike" of almost full power for a few milliseconds on pressing the PTT. This can cause damage to sensitive transverters

This is the circuit that G3UKV uses with his FT817 as a driver for his microwave transverters.

The transceiver seems to be an ideal microwave IF and he uses it both on 144MHz and 432MHz in this manner.


POWER CONTROL FOR THE FT817 TRANSCEIVER
TX ground on ACC socket
PP3 battery
+
-
68K
trimpot
50K
68K
12K
-4.5V approx
PTT link
ALC
Ground
TX ground
FT817 ACC SOCKET viewed from rear of TCVR 8 pin MIn.Din

Transverter Interfacing and other Useful Tips

Bill, W3IY

W3IY's "Rover" station

Transverter interfacing

Lot's of guys run into trouble when they try to get ready for the higher bands. Many modern-day rigs do not have built-in transverter ports, so the interfacing job can be challenging. I certainly don't know enough about most popular transceivers to be of great help but here's what I did with my IC-706MKII.

I use it as an IF on 144MHz to interface with transverters for 903 - 10368MHz. My transverters all want 2-3W of RF drive on transmit and I haven't blown any up yet! The major task of interfacing a transverter with the IC706 is that of making sure you have a method of reducing the RF output on 2 metres to a level compatible with your transverter. Most popular transverters can live with 2-3 watts of drive (e.g. the DB6NT series … editor) and the nice units from Down East Microwave can be ordered for nearly any level you would want to use. For the 2W level, I built a simple negative power supply board using the 7660 voltage inverter IC. The 706 has an external ALC input, which will limit the transmit power level when it receives a negative voltage between -5VDC and 0 VDC. I just used some perforated PC board material from Radio Slack, and took a 7805 regulator to drop the vehicle voltage down to +5VDC. Then this voltage is fed to the 7660, which turns the +5V into -5VDC, at low current. The ALC input doesn't load down this supply very much, so even though the 7660 only puts out 20mA, max, things will be OK.

Take the output from the 7660 to a 10-turn pot and run the wiper into the ALC input. Of course, you should also add some .01uF bypass capacitors, and some RF chokes, or ferrite beads, to keep things cool from RFI. I set my IC706 for a low level using the built-in power reduction feature, and then keyed up the rig on CW and set the pot for 2.5W output on 144.100MHz. If you are thinking about this carefully, you may ask “Why not just use the IC706 built-in power control?”. Well, the answer is that it produces spikes in the RF output amplitude, which can blow up your favourite transverter. I think the reason is that, internally, the 706 filters this line with a capacitor which slows down the response time. It only takes a few uSec to zap your expensive transverter however.

I recently experimented with my FT-100 to see what was required to interface it to my DEM 222 transverter, using a 28 MHz IF. I found that, even with the RF power on HF turned all the way down, there was a 20W RF power spike which occurred whenever I pressed the microphone PTT switch. It only lasted 5 milliseconds or thereabouts, but it's enough to toast a receiver mixer stage (I didn't do this, fortunately). Many guys have blown transverters by failing to realize the existence of this RF power spike. After the initial spike, the RF level was a well-behaved 3W, which was about what my transverter requires. I believe the addition of negative voltage to the FT-100 ALC input will fix this problem, enabling me to use the FT-100 with the DEM transverter. There's a good chance that the transverter would survive this spike, if I could be sure that it was switched into the transmit mode, before the spike arrived. This requires some sequencing, using a foot-switch or a delay-generator to prevent the 28 MHz transceiver from being keyed until after the transverter is keyed into transmit.

We're working on a way to do this, and still use the FT-100 microphone PTT switch. More information will be made available after some testing.

Another successful approach that can prevent RF power spikes from damaging your transverters is to disconnect the DC power from your IF exciter power amplifier. This is a proven approach for the IC706 transceiver, and it's probably applicable to the FT-100 and other radios as well. This idea is from K6LEW who spends lots of effort helping some of the K8GP rovers configure things for the contests. Owen says that it's necessary to go inside of the IC706 to do this, but that it isn't difficult to find the right place to disconnect DC power. About 100mW is still available from the RF output connector, as the RF leaks past the disabled PA stage. The beauty of this approach is that it's easy to restore normal operation when driving transverters is no longer desired (preserving resale value of the IC706). The drawback is that is disables the transceiver from being used for full power operation on the IF frequency (unless switching is added).

Another area which is dangerously ignored by many VHF and above operators is that of sequencing. Lot's of guys have gotten away without doing it but most have already, or will soon be, paying a price for this shortcut. All vhf systems that I am aware of should use sequencers!

Think about it … if you try to key up your transverter, T/R switch, power amplifier, etc, from the exciter, the RF will already be coming out of the radio before the external gear has had time to switch. Yeah, I know about the QST product reviews that show a variable delay in the CW RF output but I don't think any of these rigs provide a delay in the RF output for SSB. (The IC551D that I had many moons ago actually did have a bucket-brigade delay line inside which did perform this job … but it was mind boggling to monitor yourself in headphones with this delay happening).

RF Relays can take up to 100mSec or more to fully close, and hot switching most RF relays will damage them (now you know where many of those "bargain" RF switches at hamfests come from!). Anyway, what I am trying to say is, that to be safe, you really need an external foot-switch or toggle switch of some sort to trigger your VHF-microwave station into the transmit mode … you can't reliably use the microphone PTT contacts, unless you run them through another box of some sort.

If you really want to be crafty, you can build a microphone interface unit which will accept the microphone PTT contacts, and then drive the sequencer, which then drives your exciter to place it into the transmit mode. I like have a separate toggle switch for CW but many operators stick with the microphone PTT switch to trigger the station. Most high-performance stations still use a footswitch, to my knowledge. Anyway, you need to take your PTT source and trigger a sequencer from it. The sequencer should then

1) first close the antenna T/R switch (and the IF relays, if your transverter uses them at the RF in/IF out),
2) the sequencer should key any power amplifiers being used, and
3) key up the exciter… in my case the IC706. Most radios that I know of have a keying input line for remote transmit activation.

How long should the time delay be for these steps? It depends on your specific equipment, but typically, 50 milliseconds is long enough between each of the above steps. DEM makes excellent sequencers, although they are a bit large.

DB6NT has a nice series with various current output levels up to >10A DC current. These latter units have less flexible voltage levels, and they use P-channel MOSFETs to switch large currents.

The real beauty of the DB6NT sequencers is that they are a little larger than a postage stamp! Very nice to add inside a power amp or something.

DC Switching

Speaking of P-Channel MOSFETs, these puppies are awesome! Using them, you can implement a small high-side DC switch, using a ground-to-transmit-high-Z input. All you need is about a 22K pull up resistor between the +V input (the source), and the gate lead (control input). When you ground the gate, high-current DC is available at the drain lead. Use a fuse because, if you don't, the low Rds-ON of a good P-channel FET can produce a nice pyrotechnic display and smoke your equipment! Anything that provides a ground will trigger the largest P channel FET. Don't forget to use steering diodes, though, if you plan to switch multiple things from one set of PTT contacts.

Also, many modern rigs have a very limited keying output signal which can only sink a small amount of current...perfect for P-channel MOSFETs. The best FET I have found is the IRF4905 device from International Rectifier. These bad-boys will switch 70 Amperes!! … and they are still very high-Z input … what's not to like? They are available from Digi-Key (albeit at highly inflated prices). For a good deal on these FETs, go to an electronic supplier like Newark, Future, or Avnet where you can get them with a credit card for about $1.00 each! These things are the control element-of-choice for solving many switching problems.

The Rds-on is lower than most pieces of wire!! It's like a true solid-state relay for positive supply currents. Have fun and get those transceivers talking on the microwaves. We're waiting to work you!

Dual-Band Feeds

Dual-band feeds are really handy for roving. W5LUA, W5ZN, and others have done a great job describing these feeds in one of the Microwave Update proceedings. It's wonderful to get 2 dishes for the hassle of just one.

However, if you want to put some horsepower into that feed, BEWARE! The DEM 5.7/10GHz feed has pretty good isolation between the 5.7GHz port and the 10GHz port, when you are transmitting on 5.7GHz, but on 10 GHz, there is a lot of leakage into the 5.7GHz rig. If you are running more than 100mW, I strongly suggest that you consider a simple modification to your hardware. I am using +12V on both band's T/R relays, and I ground one side to close the relay for transmit. (this is a common way that folks use). Here's what you can do...if you also are doing a ground-to key on your RF switches. Run a wire from the ground-to-transmit terminal of the 5.7GHz relay through a diode, with the cathode away from the relay. Run this line to the control line feeding the 10GHz relay. Then, do the same thing from the control terminal of the 10GHz relay, through the diode, and over to the control circuitry of the 5.7GHz hardware. Now, Whenever you transmit on either band, you will throw both RF relays into transmit, but you will only key-up the band that you are transmitting on.

I find that I can leave the GaAsFET preamps fired up without any problems, but some guys like to switch them off during transmit. I don't think it matters, as long as you keep the preamp inputs down to 0dBm or less. (some have reported +20dBm as the threshold of pain...but you could damage the NF with less). This is a really big problem for me on 2304/3456. I have a dual-band feed for these bands, and the isolation from 3.4GHz to the 2.3 GHz port is very low...like 12dB or so. If I light off my 40W on 3456 into this puppy, I get more than 2W of juice into the 2304 preamp. Ouch!!%&*#(@__+!! Fortunately, I haven't done this but I plan to install it before fielding my 2.3/3.4GHz dish assembly again.

Good luck and protect your precious preamps!! Allegedly, the newer dual-band feeds from Directive Systems (& Down East Microwave) are better tuned to optimize isolation, but you could still get reflected RF power from nearby antennas or ground-clutter that could cause problems.

Transverter Interface for the IC746

(and other HF Transceivers)

Andy Talbot, G4JNT

The IC746 transceiver's output power is adjustable over the range 5 Watts to 100W from a front panel control, and the obvious solution is to simply turn this to minimum and use a 5 Watt rated attenuator to reduce the transmit RF to the few milliwatts needed for transverter driving. This, of course, is a recipe for disaster! Remembering to have to set the drive control to minimum each time a transverter is used means that it would be very easy to unintentionally apply 100 Watts to the attenuator, almost certainly destroying it and with the potential of damaging the transverter further up the chain. One solution would be to simply rate the attenuator at 100 Watts but that would just be plain silly!

Automatic Level Control

An ALC input to a phono socket on the back of the rig is intended primarily for use with external linear amplifiers to allow these to operate at the correct level by varying the RF input to them automatically - equivalent ALC input connections are usually available on other makes of transceiver. The handbook specifies an ALC control voltage in the range –1V to –4V and tests with a variable voltage applied to the ALC socket confirmed that the RF output on all bands could be reduced to zero with a voltage that in fact only had to vary over a relatively narrow range around minus 1.6 volts. In practice, the absolute value of this voltage will vary for different transmitters and is likely to shift with temperature but the relatively narrow range needed – which points to a high ALC loop gain – will be a common feature, ensuring constancy of output level from external linear amplifiers in spite of any gain variations over the operating bands. For interfacing to transverters, a negative voltage derived by rectifying the RF from the antenna connection directly, then feeding back to the ALC input ought to make it possible to set the wanted RF level.

Fail Safe Operation

In its simplest case, this is all that is needed for the transmit RF part of a transverter interface – a 50 ohm load with RF pick off, and diode detector across it. However, we still need to provide transmit receive switching since most transverters with low level inputs have separate transmit and receive IF ports and there is still a potential danger here in that the reduction of RF is governed by a physically separate connection. If this connector is forgotten or fails, the full 100 watts could still appear on the transverter input which, if it is now designed to accept just 40mW, could be catastrophic. The method of preventing damage from this failure mode is to add a relay at the interface input, controlled by the rectified voltage, such that if too much RF is applied the relay opens, preventing overdrive and damage. This does mean that the transceiver will now be operating into an open circuit, but its own self protection circuitry - which is always present in any self respecting rig - can now come in to play and prevent damage to the PA stage. An audible warning of this situation such as a buzzer helps to immediately identify a problem. Obviously, the relay contacts need to be normally open and the relay pulled in for normal operation, so that if RF is applied without DC power being present on the interface no RF can pass into the attenuator elements. As it is difficult to use a negative voltage to control switching circuitry when a can usefully be used here. Having the two separate level detectors for different functions is, in any case, good engineering practice as it separates the level control feedback from the overload warning circuitry.

Other Circuit Elements

A complete transverter interface will provide separate ports for transmit and receive. To do this needs transmit / receive control from the transceiver to direct the RF to the appropriate interface port; and, since this switching function is now available, provision of switched Tx and Rx voltage supplies for transverter hardware might as well be added. All transceivers provide this control output for linear amplifier switching, usually in the form of a switch closure to ground. On the IC746, the Tx/Rx connection is on a second phono socket adjacent to the ALC input – what a coincidence – but, surprisingly, not duplicated directly in the accessory socket.

Complete Transverter Interface

The circuit diagram of the complete interface is shown in Figure 2 and works as follows. The negative ALC voltage is derived from D1/D2, these are driven from a capacitive potential divider to lower the voltage across the diodes should the full 100 Watts be applied. VR1 sets the ALC level and therefore the RF power that needs to be dissipated. A 5 dB power attenuator, capable of absorbing 3 W continuously, follows a normally open relay contact in the RF input path. This attenuator is left in circuit on receive as well, but there should always be sufficient IF gain in a transverter to permit a small amount of attenuation in the return path. The second overload detector sits at the output of this 5dB attenuator and consists of D3/D4 feeding into a level
sensing circuit. This maintains the relay RLA1 pulled in, closing the RF path. The threshold voltage is set by the values of R1, R2 and ZD1. With the values shown, the relay will trip at approximately 1.5W at this point, equivalent to 4.5W from the transceiver.

The DC supplies for Rx and Tx DC power are switched alternately by two P channel MOSFETs controlled by the T/R line. The devices specified can carry 5 Amps continuously but can be directly replaced by higher (or lower) rated devices if needed. Some time ago I adopted a UHF / Microwave transverter driving standard with 144MHz IFs of 0.5 to 1 Watt on a single coax lead, with a DC voltage superimposed on this on receive only for T/R control This standard is compatible with the old IC202 144MHz SSB transceiver with its ALC pot wound to maximum. By connecting the DC coupling link and taking the output connection from point 'X', this control standard could be generated from the interface here, ensuring compatibility with all existing transverters when the IC746 is used for 144MHz.

Construction

Only the RF path is layout-critical and a suitable 80mm x 50mm PCB layout is shown in Figure 1.

This employs mostly surface mount components and wire ended resistors for the higher power attenuator elements. No precise component details or list is shown as most of the components may be substituted by ones of a broadly similar kind.

Depending on the drive frequency chosen, the value of the coupling capacitors and DC chokes may need to be altered. The values shown are optimum for the 5 to 50 MHz range; for a 144MHz drive frequency the 10nF capacitors could be reduced to 1000 pF and the DC chokes to 220nH. VR1 is adjusted to set a level of no more than 5W carrier from the transceiver to avoid cooking the 5dB attenuator. If this level trips the overload circuitry, either reduce the value of R2 slightly or operate at a lower interface level. A suitable point might be to trip at the 7 - 8 watt region as modern metal film resistors should be able to survive this sort of abuse for short periods. By connecting a DC operated piezo sounder at the point Z, audible warning will be given if the interface encounters an overload situation.

Summary

The technique shown here, using a commercial transceiver's ALC input to control the level supplied to a transverter and backed up with additional independent overload detection circuitry, provides a safe interface that will prevent damage to transverter equipment if connections are left off or fail. The only residual danger that can still occur is if the transmitter cannot operate safely into an open circuit - but this is only likely with ancient valve PA rigs and these often have a low level output (with the PA heaters switched off) for driving transverters anyway.........

FIG. 1: PC BOARD LAYOUT – G4JNT INTERFACE
The pc board measures 80mm x 50mm

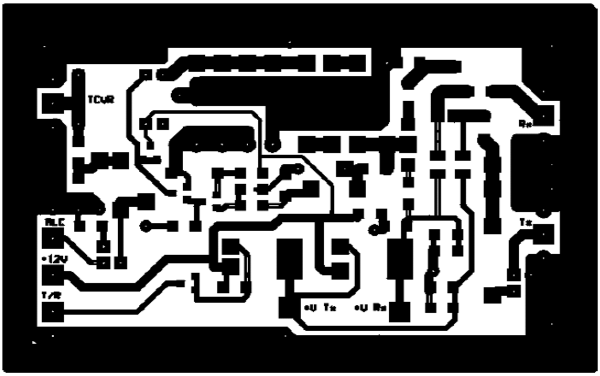


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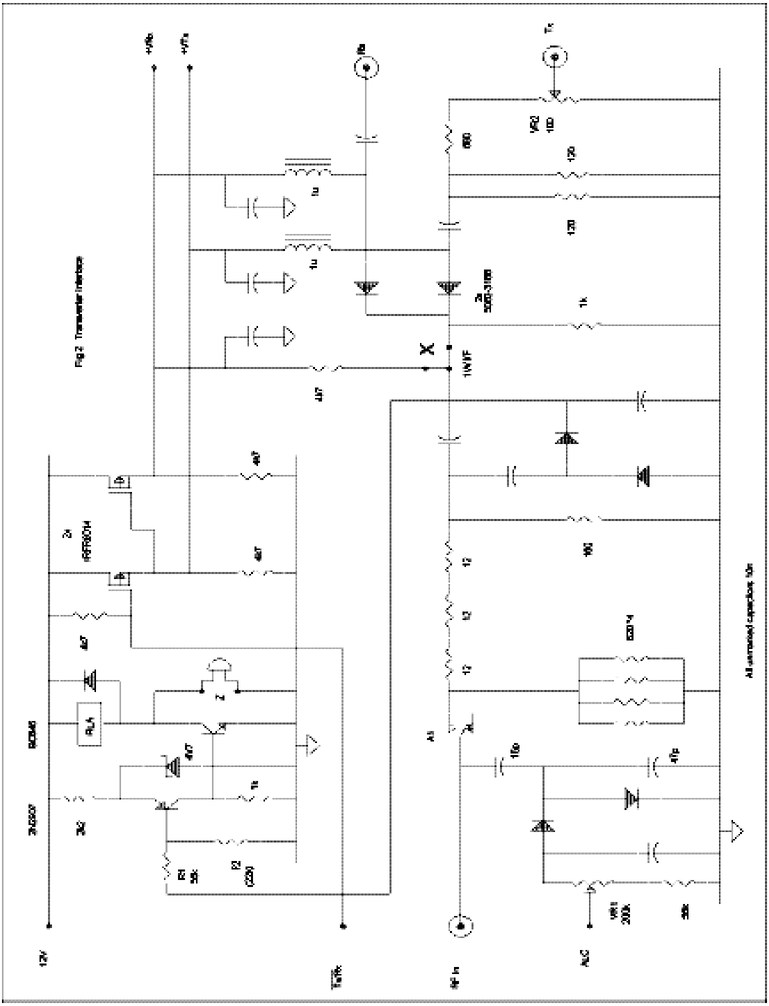
FIG. 2: CIRCUIT DIAGRAM – G4JNT INTERFACE



Modifying the FT736R for Microwave Transverter Switching

Keith, GW3TKH

Until now, I have used a FT290R as the 2m IF for my transverters at home and portable.

In transmit mode, this transceiver has a +9V phantom voltage on the RF at the antenna socket which is used to switch the transverter.

I recently acquired an old FT736R for use as a microwave IF in the home shack and investigated modifying it to provide a similar switching voltage. This turned out to be quite easy, as the FT736R has a switchable receive pre-amp supply applied to the antenna socket and is selectable internally with 'U' links. It is normally deselected (see the handbook), which is the condition we need, as some of the wiring is used in this modification.

Proceed as follows:

1. Remove the top cover, taking care with the speaker lead.
2. Referring to the 144MHz PA UNIT circuit diagram, remove the BLUE wire from feed through insulator 'PRE'. Insulate and tie back the wire.
3. Take a 10k 0.25W resistor, fit a small ferrite bead on each lead, solder between feed through insulators 'TX9V' and 'PRE'. This puts +9V on transmit at the ANT socket. (If a dc short circuit is applied to the ANT socket, i.e. a 2m antenna, the max current drawn is <1mA, so no damage will occur.)
4. If necessary, the power is set up to the required maximum to drive your transverter. (see adjustment details by GM4PLM on www.mods.dk)
5. Replace the top cover, not forgetting to reconnect the speaker lead.

This modification is easily reversed if required and/or can also be applied to the 70cm PA. I have been using the FT736R modified as above, set to a maximum drive level of 500mW, to work with microwave transverters by DB6NT and SSB Electronics. No adverse effects have been noted!

Chapter 9

Power Supplies

Switching Regulator for Light Duty Applications

John Hazell, G8ACE

There was a recent need to provide a regulated 12v dc supply from a 12v battery. Using a standard 78L regulator there would be insufficient supply voltage available for regulation. The solution adopted was to use a TL497 switching regulator to provide a higher supply voltage. Whilst this regulator can provide a regulated output directly, in the particular application the output was set to 15v to feed a 78L regulator. This is a 14 pin IC and includes an on board rectifier diode. The total component count for the voltage booster is therefore quite small.

The TL497 IC (available cheaply from Farnell) can be used as a step up, step down or inverting supply circuit.

Editor's comments ....

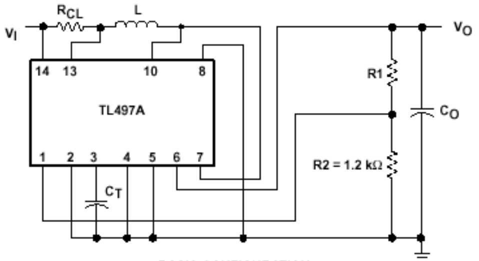
This design can be used for microwave changeover relay supplies as it can easily supply the 24V or so required by the ubiquitous Transco and Dynatech types.

The following component values on a prototype produced an output of 24.2V, for a minimum of input 8.5V, thus giving an enormous amount of headroom for a flagging 12V battery!

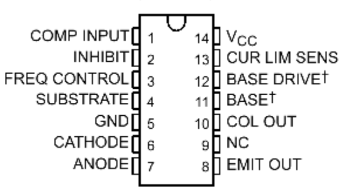
R1: 1K2, R2: 23K (made up of 10K and 13K in series) CT: 4n7, CO: 0.33uF. 470uF electrolytics were fixed across input and output supplies.

According to the data sheet, the device is good for 500mA maximum but the higher the voltage out the lower the available current. However the 24V supply should be good for several hundred mA, more than enough for a relay supply.

The TL497A is a versatile chip and can be used in a step up, step down or inverting supply circuit. Remember that it is switching at 50kHz or so and thus will require careful filtering.



BASIC CONFIGURATION
(Peak Switching Current = $I_{(PK)}$ < 500 mA)



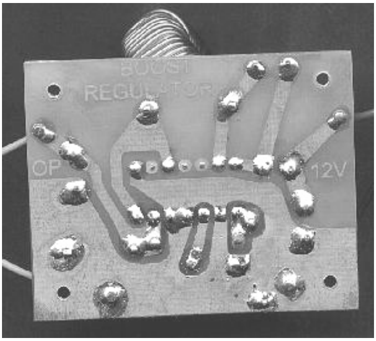
DESIGN EQUATIONS

- $I_{(PK)} = 2\, I_O \max \left[\frac{V_O}{V_I}\right]$
- $L\ (\mu H) = \frac{V_I}{I_{(PK)}} t_{on}\ (\mu s)$

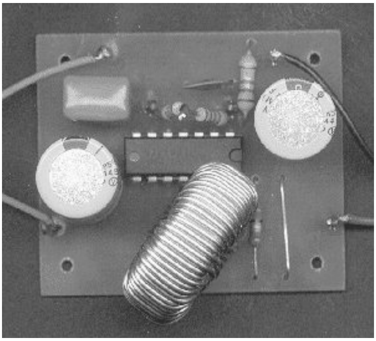
Choose L (50 to 500 μH), calculate t_{on} (25 to 150 μs)

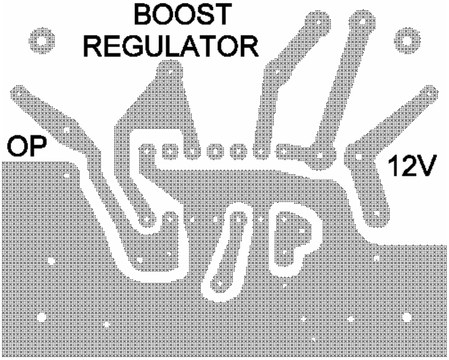
- $C_T(pF) \approx 12\ t_{on}\ (\mu s)$
- $R1 = (V_O - 1.2)\ k\Omega$
- $R_{CL} = \frac{0.5\ V}{I_{(PK)}}$
- $C_O\ (\mu F) \approx t_{on}(\mu s) \frac{\left[\frac{V_I}{V_O}\ I_{(PK)} + I_O\right]}{V_{ripple}\ (PK)}$

The TL497A switching regulator is a cheap alternative to the LM2577T described in the Microwave Newsletter some time ago, especially when no more than a few hundred milliamps of output current are required. For supplies in excess of 0.5A, the LM2577T is to be recommended.



Above and below: the printed circuit board used in the proto-





13.8V from your Car Battery
Modifying the Vanson SDR-120W

Chris, G3WIE

When I started on microwaves a year ago, I had to rediscover /P operating. Given that I was only running QRP (1 watt) from a borrowed 144MHz - 10GHz transverter, a morning's operation from the cigar lighter socket in my car seemed quite possible but I soon discovered that the transverter required a stable 13.6 volts, not a gradually reducing 12.6 volts and I've seen others describing the same problem in Scatterpoint through 2005.

Looking around the local Maplin shop, I found a 120 watt DC-DC converter which supplied a range of laptop computer voltages: 15-16-18-19-20-22-24 volts selling for £30 (stock number: L40BB). Gambling that I'd be able to modify this to reduce the lowest voltage to 13.8 volts, I bought one. It turns out that you have to change just a single resistor. This article describes how I did the modification.

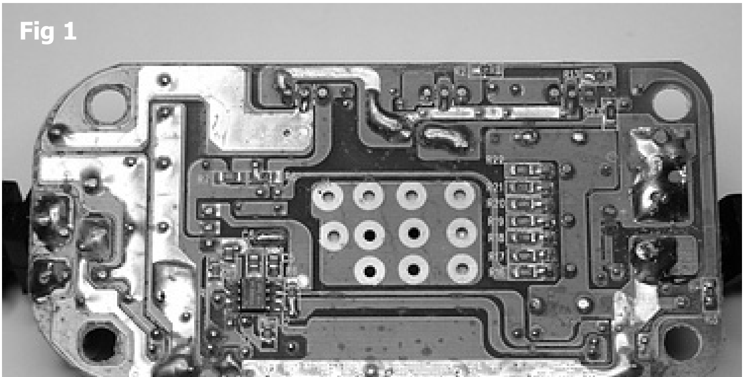
You will need:

- a 5mm flat-blade screwdriver which you are prepared to modify (to remove the tamper-proof screws holding the case together)
- a set of needle files to do the screwdriver modification
- four 12 mm x 3mm self-tapping screws to replace the above!
- a temporarily-modified soldering iron to remove an SMD resistor
- a replacement SMD resistor or small wire-ended one

Dismantling the unit:

The case is held together by four tamper-proof self-tapping screws. To remove these, modify a 5mm flat-blade screwdriver by filing a rectangular notch 1.5 mm wide and 1.5 mm deep in the centre of the blade. Now you

Fig 1



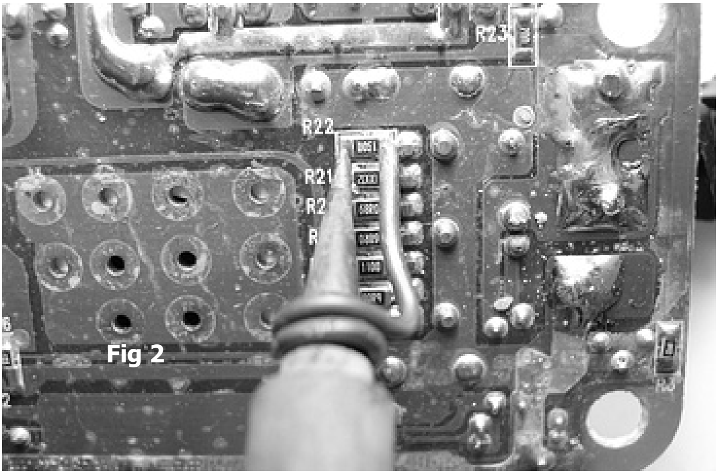


Fig 2

can remove the screws without wrecking your small pliers!
Set the voltage selector switch to the 15 volt position, remove the silver-coloured top of the case, then lift the PSU assembly out of the other half of the case. The brass screen fitted underneath the circuit board may jam on the plastic pillars, but if you bends the screen to get access it can be straightened afterwards Slide the screen and its insulation out from under the three clips on the heatsink. The clips may jump off – replace them again as they hold semiconductors on to the heatsink.

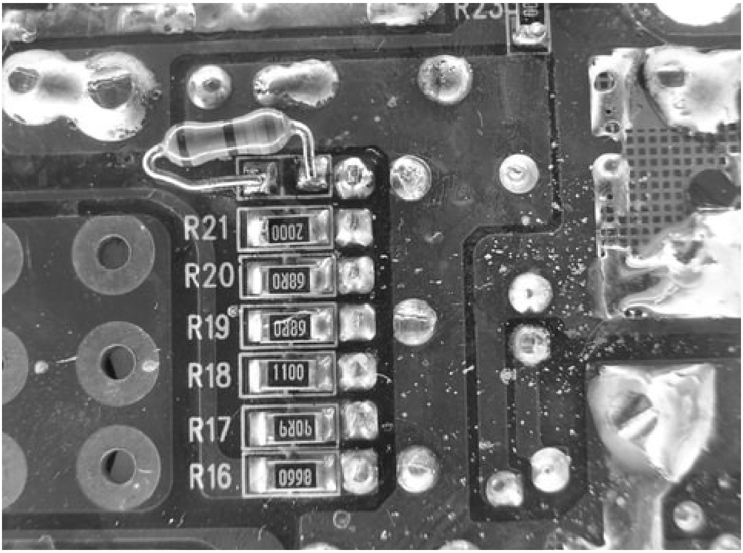
Fig 1 (previous page) shows a picture of the unit with the shield removed.

Modifying the unit:
On the underside of the unit, towards the end nearest the input
cable, you will see a row of surface mount resistors labelled R16 to R22, connected in series. The voltage selector switch connects the junctions of these resistors to ground – increasing the total resistance from the top of R16 to ground produces a lower output voltage, so increasing the value of R22 will reduce the voltage at the 15V setting without affecting the higher voltage outputs.
First, note the value of R22 (130 ohms in my unit), then remove the resistor. I don't have two small soldering irons, so I modified my one and only iron by wrapping two turns of 16swg tinned copper wire round the bit and bending it so I have a double-tipped iron. File a point on the wire and you're ready to go.

Fig 2 shows the idea - it's a posed photograph but this technique does work. Usually the resistor will stick to the tips of the iron but be ready with a pair of pointed tweezers in case it doesn't.

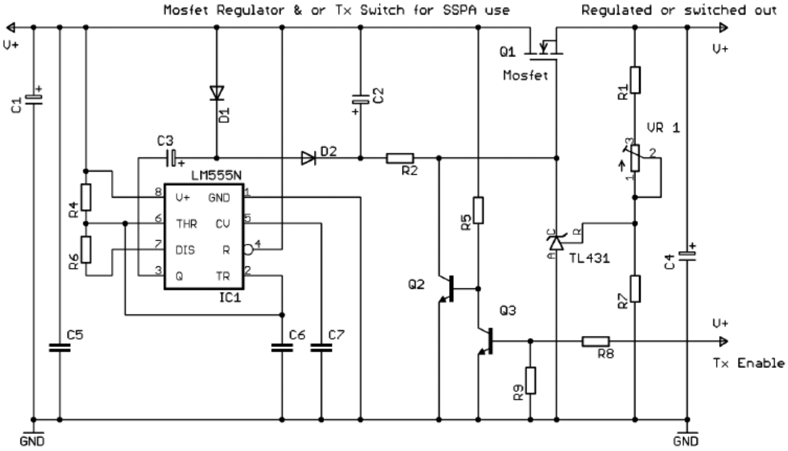
Now fit a replacement resistor. **Fig 3,** on the next page, shows how I fitted mine.
Plug the converter into your car and check the output. In my case 390 ohms at R22 produced an output voltage of 13.8volts off load, but I fitted a 330 ohm to provide a bit extra to compensate for the voltage drop along cables.
Finally, reassemble the unit, ensuring that the clips that held the screen are holding the semiconductors against the heat sink.

In use:
I use the converter to power my entire station which draws about 3 to 4 amps on transmit and 1 amp on receive. The converter runs cold even on a hot day. I have managed to short circuit the output at least twice. Each time, the 16 amp fuse I had fitted in the battery supply blew but the converter survived!



A MOSFET Regulator and/or Switch

John Hazell, G8ACE



Circuits for MOSFET regulators seem rather sparse and so this circuit was built up referring to a just a few circuit suggestions found during a web trawl. The need for the regulator arose as a low voltage drop, high current regulator was required to drive an HP 9cm 15w SSPA unit. Tests showed the SSPA gave full power from a supply of 11.7v upwards. However the upper safe voltage operating voltage was unknown so it seemed a good idea to utilise a low drop regulator to provide +12v even when running from a +13.6v supply or fully charged batteries to ensure a long PA life. The circuit below also serves for the high current part of the Tx sequencer.

An N type MOSFET is employed. These are fairly inexpensive and the criteria for selection is: Cost, Current handling, On resistance, preferably less than 0.01ohm, plus adequate voltage rating. The STP70NF3LL from Farnell was used but typically for MOSFETs its already no longer available. Device availability lasts just for a short span of time so refer to its data as a guide for other suitable selections. A low on resistance device will drop only tens of millivolts at several amps in the switch arrangement. Of course the package needs to be at least TO220 to handle the dissipation especially when used as a 9.6v regulator for a module such as the Ionica SSPA. The down side of the N MOSFET regulator is it needs a gate supply higher than the incoming supply, a minimum of 16v is a good target for an incoming nominal 12v battery voltage.

The 555 IC is used with a voltage doubler for the gate supply. Download the 555 data and refer to the 50:50 square wave oscillator. Note R6 should be less than one half the value of R4 to achieve the square wave for efficient rectification. Values used were R4 47K, R6 18K and C6 C7, 0.01nF. These values produce a low frequency so that there will be minimal interference produced onto the dc supply. C2,C3 need to be adequately sized according to the frequency used. The gate supply can also be derived from your +24v antenna relay supply of course.

The regulated output from the MOSFET is achieved by using the inexpensive TL431 shunt regulator. Again download the pdf for reference. The device current should be not less than 1mA so R2 must be selected accordingly depending how the gate supply is derived. 3K3 was used from a 26v relay supply. The reference voltage for the TL431 is 2.5v between reference input and anode (ground). R7 is non critical so using say 1K then the value of VR1 say 500W the value needed for R1 can be calculated according to the output voltage needed. Two transistors are added such that the regulator can be keyed on for Tx with a +ve supply connected to R8. By sourcing the +ve voltage to R8 and Q3 from a sequencer the regulator eliminates the need for a separate MOSFET switch in the PA supply. R9 can be around 100K. If the regulating aspects are not required then the TL431 can be omitted so the MOSFET device simply acts as a low voltage drop switch. Powering an SSPA using the MOSFET in the switch arrangement the supply voltage drop was typically 60mV

Extensions to the MOSFET Regulator

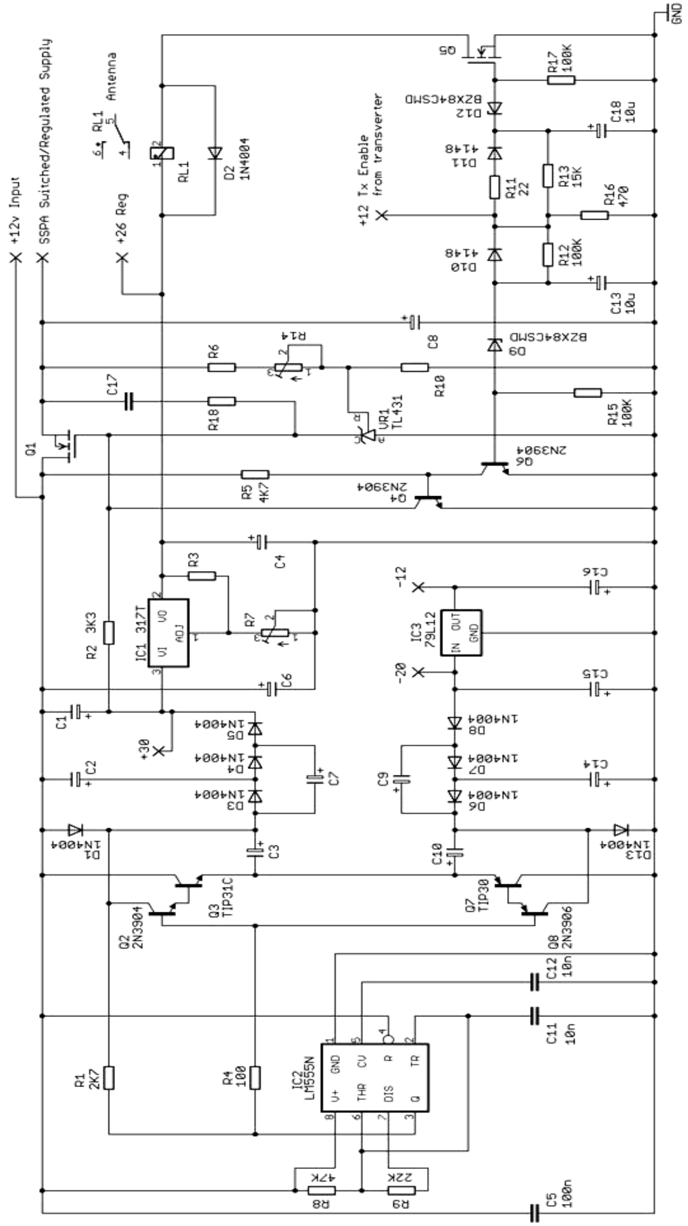
John Hazell, G8ACE

The **MOSFET Regulator and/or Switch** (previous article), for use with an SSPA, can be extended to include the other supplies needed by a home brew microwave transmitter using, for example, an Ionica 9cm Tx module. This is 'low tech' circuitry for which most parts can be found in the average 'cum handy' box.

Additional functions added here to the original circuit are:
Relay supply, -12V Bias supply
and Sequencer. These circuits have, of course, been published before in some form or other, but perhaps not with these variations. The relay supplies formerly published in newsletter pages either use a switching IC regulator or voltage doubler. It's difficult to achieve more than 20 volts under load from a diode voltage doubler using a 12V battery supply. Consideration of the transistor and diode voltage drops will confirm this. It was thought rather risky to rely on such a low voltage to reliably operate a 24-28V antenna relay carrying watts of RF power.

The IC switching regulator style relay source whilst being efficient does produce problems with the 50kHz signal creeping into unwanted areas of the rig unless very good supply filtering is used hence not being used here. In the previous MOSFET Regulator/switch circuit, the 555 IC is used to supply the higher than incoming rail voltage needed for the gate of the MOSFET. This 555 is now additionally employed for the relay and bias supplies, as commonly seen before, but this time with two changes. The 24-28V is generated by a tripler rather than doubler. Complimentary transistors drive the multiplier diodes driven by bootstrapped emitter followers Q2, Q8. Referring to the circuit note the collectors of these emitter followers are connected to the first stage of the dual polarity voltage multipliers. The increased voltage on the followers enables the drive into the complimentary pair to be maintained as the bases are driven down to 0v and up to near the 12V rail. This increases the efficiency by a few percent as does the resistor R1 back to pin 3 on the 555. The value used for R1 must not be to low otherwise some unwanted oscillation takes place. Common 1N400* series diodes are used in the multiplier along with low ESR electrolytics all 25V except for the post regulator C4 which should be 40V. The multiplier supply is hung from the incoming supply rail hence the requirement for only 25V capacitors. C6 must not be omitted in this arrangement. Around +31V is achieved from the multiplier which is connected via a 317 to the relay and sequencer.

A standard N MOSFET is used for the antenna relay switching. Several ohms on resistance can be tolerated for this device as it is switching at most a few hundred mA. Poor voltage stability of the voltage multiplier output under load would indicate capacitors with an undesirably high ESR. The negative multiplier supply achieves -20V which is regulated by a 7912 regulator for the –12V bias supply. Sequencing is achieved by employing the same circuit technique as the DB6NT MOSFET sequencer. When +12V is applied to the Tx enable point C18 charges rapidly through D11 and R11 such that the antenna MOSFET switch operates almost instantly. The PA supply MOSFET switching on is delayed however by the time C13 takes to charge through R12, D10 being none conducting. When the +12V Tx enable is removed the PA MOSFET switches off first as C13 discharges rapidly through D10 and R16 however the antenna relay is held on until C18 has discharged slowly through R13 and R16, D11 now being none conducting. Circuit values are shown where the values are standard. Unmarked circuit component values can be calculated according to requirements. Sections of the circuitry can be omitted when not required. As before the main MOSFET regulator can be used as a PA supply switch where the PA requires the same voltage as the incoming dc supply or as a switched regulator where a lower voltage is required. Output voltage is set with R14. The resistor capacitor C17, R18 have not been used to date. An adequate heat sink is required for the regulator MOSFET when a PA voltage such as 9.6V is required from the incoming 12V. Always refer to the pdf data files readily available on the web for the various semiconductor devices used.


+12v Input
SSPA Switched/Regulated Supply
+26 Reg
RL1
6
4
5
Antenna
RL1
D2 1N4004
GND
Q5
R17 100K
D12 BZX84CSMD
C18 10u
D11 4148
R11 22
R13 15K
+12 Tx Enable from transverter
R16 470
D10 4148
R12 100K
C13 10u
C8
D9 BZX84CSMD
R6
R14
R10
C17
R18
VR1 TL431
R15 100K
Q1
Q6 2N3904
R5 4K7
Q4 2N3904
R3
C4
IC1 317T
VI
VO
ADJ
R7
R2 3K3
C6
-12
C16
IC3 79L12
IN
OUT
GND
-20
C15
C1
+30
D5 1N4004
D4 1N4004
D3 1N4004
C2
C7
C9
D8 1N4004
D7 1N4004
D6 1N4004
C14
D1 1N4004
C3
C10
D13 1N4004
Q3 TIP31C
Q2 2N3904
Q7 TIP30
Q8 2N3906
C12 10n
C11 10n
IC2 LM555N
V+
THR
DIS
Q
GND
CV
R
TR
R1 2K7
R4 100
R8 47K
R9 22K
C5 100n

Chapter 10

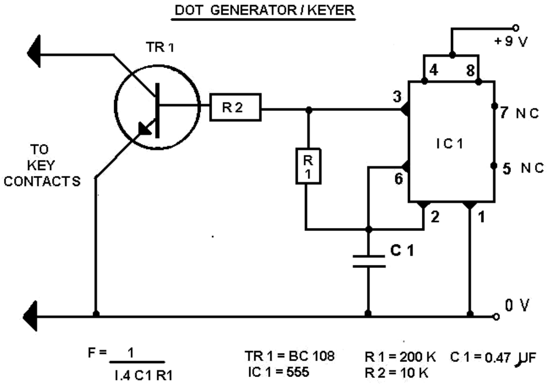
Keyers and Control

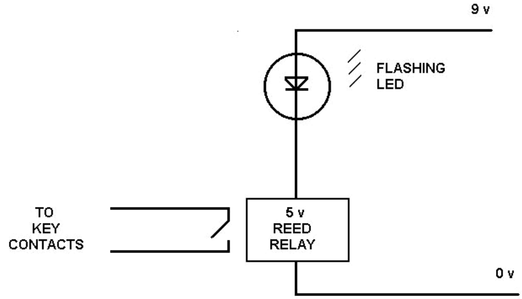
Simple Dot Generator/Keyer

Steve Cooke, G1MPW

Having decided that some sort of keyer was needed to pulse the carrier on 10GHz , I came across the basic astable circuit and added the switching transistor to key the carrier on and off. The switching rate is given by F = 1/1.4 C1 R1. With the values given (200k and 0.47uF) it produces rapid dots.

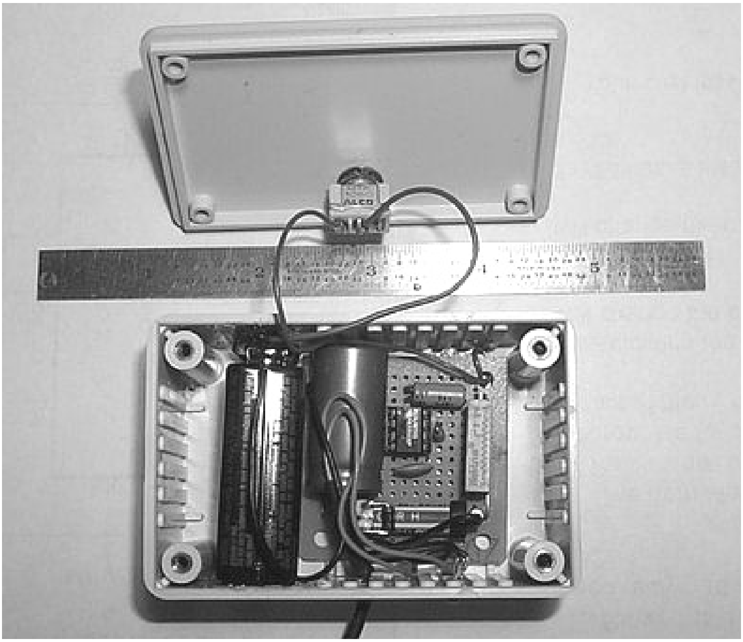
The " simple circuit " was a bit of an after thought but seems to work fine but at a slower rate of about 2Hz. On the FT290 it is necessary to close the PTT switch as well as the key contacts to TX a carrier in the CW mode . I wanted to use a non-locking push button switch but Double Pole varieties don't seem to be available so I used a small reed relay connected across the 555 timer + and - to provide the extra contacts needed for the PTT.





The Dasher …
an automatic CW dash generator

Fred Miller, WO2P

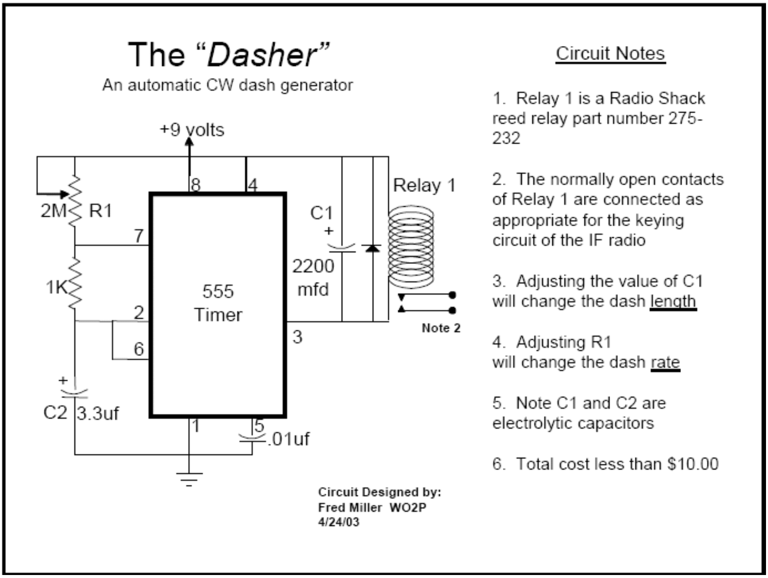


The "Dasher" was designed and built within two week of our recent 24 gigahertz Marathon. It was built purely out of a need for self-preservation. Standing on a cold windy hilltop and sending (not so well formed) dashes for 5 to 10 minutes at a time with a straight key was, to say the least, uncomfortable. I refrained from using my home station keyer because of the added weight and bulk to an already pretty hefty "portable" station. I wanted something small, light and self contained.

Version one was based on a blinking led" from Radio Shack. I demoed it at last month's club meeting. It functioned fairly well and was really inexpensive but the dash rate seemed just too slow. The blink rate was preset by the chip internal to the LED. With no way to adjust it, it was back to the drawing board.

The current version seen here is based on a 555 timer. I experimented with various resistor and capacitor values until I arrived at what I think is a good balance. When operational pin 3 will generate regular low going pulses based on the setting of R1. The low going pulse charges C1 and energies Relay 1 closing the normally open contacts and activating the keyed circuit in the IF radio. When pin 3 goes high C1 discharges through Relay one and delays its release. The length of the Relay 1 release delay is determined by the value of C1. The larger the value of C1 the longer Relay 1 will remain operated. A short period of time later the cycle repeats creating dashes.

Since my IF radio is an ICOM IC-260 that requires hard keying, I connected Relay 1's normally open contacts to a 2 wire cable with an 1/8″ plug on the other end. I inserted a 1/8‟ "Y splitter" into the IF radio CW key jack and then my straight key plug and dasher plug into the splitter. The straight key and the Dasher are in parallel. When I need to send dashes I just flip the toggle switch on the dasher. When my signal is found I switch off the Dasher and finish the QSO with the straight key. The entire circuit including a 9 volt battery fits into a small 3.5″ x 2.5″ x 1.5″ plastic case that is bungie-strapped to the side of my portable microwave system. If you are a "hill-top Microwaver" and looking for a very inexpensive one evening project give this one a try. You will thank yourself the next time you are out on a cold windy day.



The "*Dasher*"

An automatic CW dash generator

Circuit Notes

1. Relay 1 is a Radio Shack reed relay part number 275-232
2. The normally open contacts of Relay 1 are connected as appropriate for the keying circuit of the IF radio
3. Adjusting the value of C1 will change the dash length
4. Adjusting R1 will change the dash rate
5. Note C1 and C2 are electrolytic capacitors
6. Total cost less than $10.00

Circuit Designed by:
Fred Miller WO2P
4/24/03

On-Off Keying for Microwave Beacons

Andy Talbot, G4JNT

The normal frequency shift keying applied to beacons is no good if the intention of the beacon is to provide a very high stability frequency standard and in this case it becomes necessary to on-off key the RF note 1. Some time ago I built a simple personal beacon based around a DDK004 modified for external source and WDG001. On –off keying was done with the aid of a PNP transistor keying the positive supply to the GaAs FET devices in the WDG001 refer to Fig 1.

Recently, whilst waiting for the new site clearance for GB3SCX to go through, I wanted to install the beacon head unit, with my high stability source, as a personal beacon from the work QTH on Portsdown Hill. GB3SCX is built in two parts. The lower section contains a DDK004 source, modified to apply +12V to the feeder for powering the beacon up the coax. Conventional FSK is used for keying. The head unit contains the final WDG001 X4 stage and 120mW power amplifier. An additional stage at the input contains a MSA02 modamp and DC take off. The modamp is there to permit a high feeder loss for the 2.6 GHz signal, and means the beacon will give full output for a drive as low as –2dBm. This allows a feeder loss of up to 12dB, with overdrive not being a problem as the MSA02 just saturates.

To on-off key this system is not so straightforward, as the WDG001 is now at the head unit and I certainly didn't want to install extra cabling just for the keyer. The solution was to key the 2.6GHz RF feed from the base unit. Initially I just made up another modamp gain stage using a MAR-2, and keyed the supply to this with a PNP transistor.

The circuit is that to the right of the shunt diode shown in Figure 2, the modamp being fed directly with RF. This stage managed a gain of 5dB at 2.6GHz, and saturated at 6dBm output level, enough to drive the head unit with up to around 6dB feeder loss. Keying the supply reduced the saturated output by 16dB, but when driven by the 8dBm available from the DDK004 multiplier still gave an output level of –10dBm.

Remember, the 'SCX beacon would work with a drive level of –2dBm. Applying –10dBm resulted in an output just 10dB down – not ideal for CW keying and this would certainly change over temperature (not tested though). I replaced the MAR-2 device with the higher power MSA04 and achieved a gain of around 4dB, but saturating at a much higher power. Furthermore, the amplitude change by keying the supply was now closer to 20dB., but output when driven by the DDK004 was now 11dBm – far to much for driving the head unit and still giving something like -9dB key up drive level. unless a considerable feeder loss could be added.

Adding a 6dB attenuator pad between the '004 and amplifier stage resulted in levels of 7dBm/ -13dBm which, when connected directly to the head unit, produced an amplitude change of 25dB – a lot more acceptable. By inserting a longer feeder, with around 4dB loss the key up output level could be reduced by well over 30dB – even better.

This hit and miss method of selecting drive levels was adequate for the temporary installation, but for a permanent on-off keyed beacon, some means is needed to reduce the

RF by a lot more than 20dB if sufficient margin is to be allowed for arbitrary feeder losses.

A figure of 40dB would be ideal so that amplifiers can be operated saturated and still give sufficient change in amplitude for at least 30dB final keying levels. Keying of any of the DDK004 multiplier would probably work but was not attempted; since this scheme could be used with the conventional 004 with integral oscillator any attempt to play about here would probably cause chirp. Using two modamps with their supplies switched would allow 40dB level change provided no saturation was occurring, but with a ridiculously high overall gain would need considerable pre-attenuation.

The ideal solution, is to combine modamp supply keying with a shunt diode.

See Figure 2 for the proposed – but not yet tested - circuit of the final keyer.

Note 1

One solution allowing FSK keying while still maintaining a high stability, at least on one of the tones, is to use frequency exchange keying. Here, two completely separate oscillators are used corresponding to mark and space tones, and the outputs switched alternately to the multiplier chain.

One can be made the high stability source, and the other just used for the keying. Frequency, and more importantly phase, information is preserved allowing ultra narrow band DSP based monitoring.

Something not possible if phase coherence is lost between each mark-space transition.

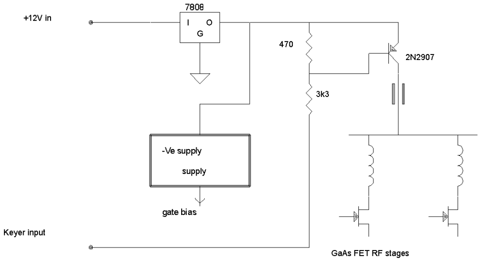


Figure 1
Amplitude keying
the WDG001 multiplier

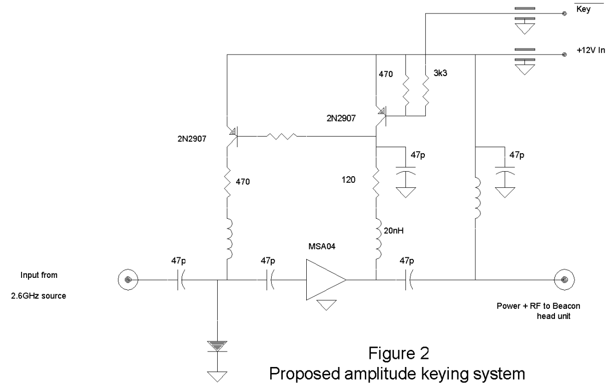


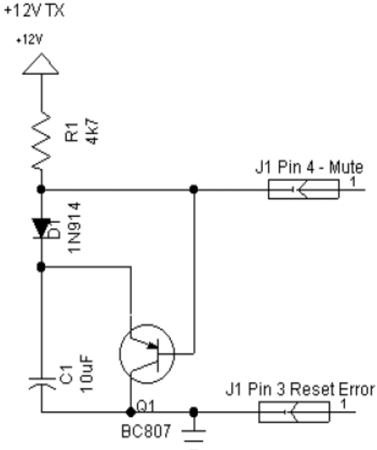
Figure 2
Proposed amplitude keying system

Mute sequencer circuit for the Ionica PA

John Quarmby G3XDY

The sequencer circuit is connected to a low impedance source of +12V on transmit which also operates the antenna changeover relay. This 12V source can be either open circuit or resistively tied to ground when on receive. When going from receive to transmit Tantalum capacitor C1 charges via R1 and D1 with a time constant of a few tens of ms, and the PA mute circuit goes to transmit when the voltage on pin 4 of the connector passes +2.5V. There is a 3k3 resistor to ground in the PA mute circuit which limits the voltage on pin 4 to +5V when the capacitor is fully charged. This allows the antenna relay to fully operate before any RF is applied. When going from transmit to receive, the 12V disappears from R1 and the internal 3k3 resistor pulls down pin 4 towards ground. This switches on Q1 which rapidly discharges C1 so that the PA is muted within a few hundred microseconds, well before the antenna relay starts to release. Flashing" the PTT will not cause problems as Q1 discharges C1 fast enough to maintain correct operation under all circumstances. The Reset Error output from the PA is grounded to stop it feeding 12V into the Mute circuit through a diode on the PA board. The device driving the Reset Error output is an open collector comparator so there are no ill effects from connecting this to ground.

The original was laid out on a tiny piece of PCB using surface mounted components, but almost any high gain PNP switching transistor should do the job if leaded components are preferred.



Quiet Fan Controller
Speed Varies with Temperature

Paul Wade, W1GHZ © 2005

Many modern radios and other electronic devices rely on muffin fans for cooling. These can be loud and annoying. Some run continuously, while others cycle on and off, either when needed or just on transmit. In some radios, the fan cycling results in a small frequency shift as the oscillator is heated and cooled. Wouldn't it be preferable to have a fan with a variable speed, responding to cooling needs?

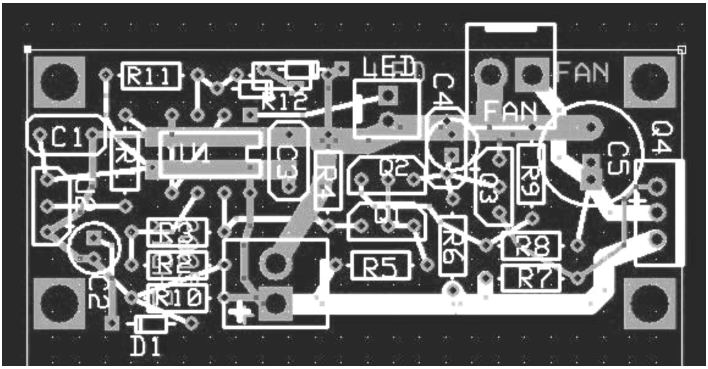
I've thought so for a long time, but never got around to doing something about it. Recently, I decided it was time. I figured this was an obviously useful thing, so there would be lots of circuits available on the web. The only things I could find were microprocessor circuits, many of them relying on fancy fans with internal tachometers – none of those in my junk box. Also, the microprocessor controls the speed by turning the fan on and off rapidly; some of the notes suggested that the results are audible. Most muffin fans use DC brushless motors, so the speed is easily controlled by varying the motor voltage. 12-volt fans are convenient and readily available. Also, there are a number of inexpensive temperature-sensing ICs available. What we need is a simple circuit to vary the fan voltage in proportion to temperature – basically, an amplifier. A couple of op amps should do the job. I sketched out some circuits and simulated them with the free witcherCAD III software from www.linear.com.

None of them worked satisfactorily, so I called the op amp guru, Byron, N1EKV. He agreed that it sounded simple and would look into it. He soon called back to say it wasn't as simple as it sounded because of some choices I had made: to keep one end of the temperature sensor and one end of the fan grounded, and to drive it with a power FET for minimum voltage drop at full speed. The result is that the sensor is referenced to ground, but the FET is referenced to the positive voltage. The final complication is that there is a huge gain in the circuit due to the transconductance of the FET, about 2.5 Siemens (in tube terms, this is 2,500,000 µmhos – a typical tube is 5000) or more. To make things worse, the FET is operating in a non-linear region, and having non-linear elements inside a feedback loop is never a good idea.

I went back to engineering basics: find a circuit to steal. One of the microprocessor fan controls used an interesting circuit to drive a FET and shift the reference from ground to high side. The circuit, in the area of Q1 and Q2 in the schematic, looks like the Widlar current mirror used in many integrated circuits. I added this circuit plus the PNP emitter follower, Q3, and fiddled with the resistances to get it going. Then I consulted Byron again and added capacitors C4 and C5 to stabilize things.

Computer simulations are only as good as the models, and don't always fully model reality, so I built up a breadboard on a piece of perforated board. It actually works – and it definitely oscillates without C4!

Now it felt safe to make a printed-circuit board, to make it reproducible and robust enough for portable equipment – perforated board wonders seem to fall apart bouncing around in

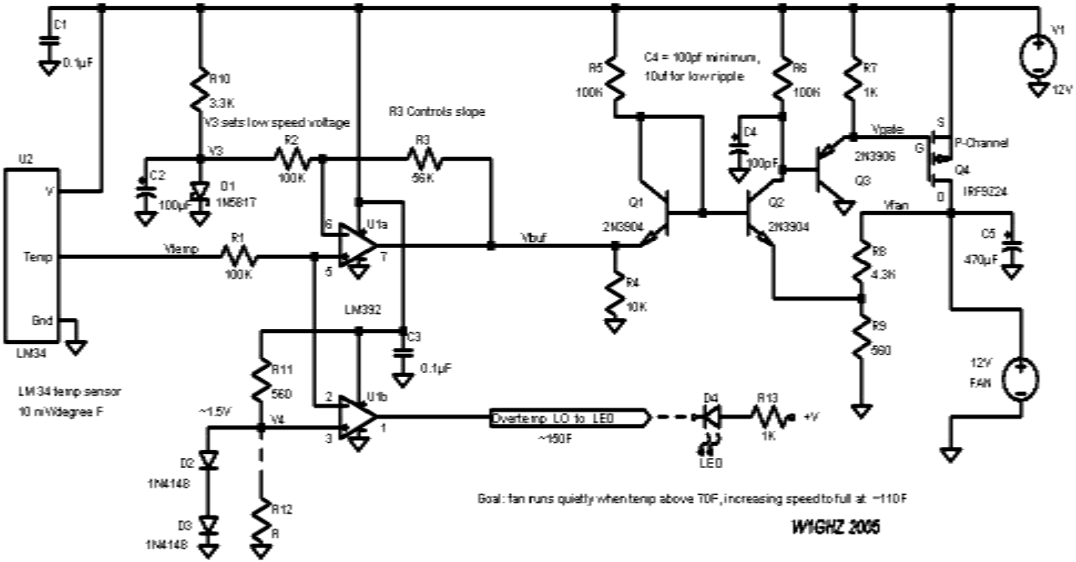


the back of the truck. Since op amps come in pairs and quads, I tried to think of something to do with the other half. The best use I could come up with is an over-temperature alert, but that really requires a comparator rather than an op amp. But there is one IC available with one of each, the LM392. The comparator uses the output of the same temperature sensor to provide an alert at some higher temperature. The output goes low at the desired temperature to turn on whatever: an LED, a sound, a relay to shut down the amplifier, or a jolt to the operator's chair. The noisemaker from a defunct smoke alarm might be interesting, but a blinking LED seems adequate to remind me not to talk so much. I used the free software from ExpressPCB (http://www.expresspcb.com) to layout the board shown in Figure 2 and placed a Miniboard order: three boards in four days for $59. Four days later, the boards arrived, I put one together and sparked it up. After I added one resistor that somehow was left out of the layout (Figure 2 includes the correction), it works fine. The fan purrs away at room temperature and speeds up as the temperature sensor is heated up.

Figure 3 is a photo of the completed controller, with the LM34 temperature sensor at the left edge of the board, not yet attached to a heat source. It could also be soldered to the other side of the board, if the intent were to mount this board on the heat sink. With the resistor values in the schematic, the fan gets about 9.5 volts at room temperature and gets up to full speed with full voltage at about 105°F. We will use Fahrenheit since the LM34 temperature sensor output is in Fahrenheit: 10mV per degree F, so the output at 70°F is 700 mV and at 105°F is 1.05 Volts. The slow speed is set by voltage V3, controlled by resistor R10; decreasing R10 increases the current through Schottky diode D1, which increases the voltage drop of the diode and increases V3. The temperature at which the fan reaches full speed is controlled by resistor R3; increasing R3 makes the fan reach full speed at a lower temperature. Note that we can monitor the temperature directly by measuring voltage Vtemp at the LM34 output, as 10mV per degree F.

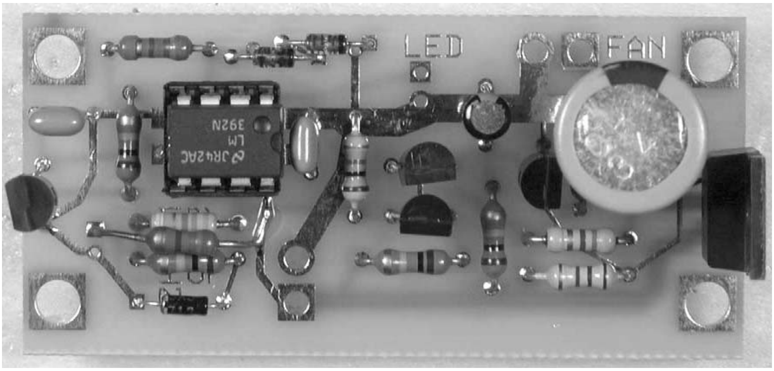
The over-temperature setting for the comparator is similarly set by voltage V4, the voltage drop through silicon diodes D2 and D3. Decreasing resistor R11 increases the current through the diodes, increasing the voltage drop and thus raising the temperature setting. With the value shown for R11, V4 is about 1.50 volts, so the over-temperature alarm is at about 150°F. If a much different temperature setting is desired, R12 could

Temperature-proportional Fan Speed Controller



be used instead of the diodes, but the temperature setting would vary with the supply voltage.

The temperature sensor U2, the LM34, should be in contact with the heat sink or surface being cooled by the fan. Either attach the flat side of U2 directly to the heat sink with Super Glue, or use a dab of heat sink compound and clamp it on. A heat sink takes some time to heat and cool, so the fan will not change speed instantaneously, but will speed up as the heat sink heats. More important, it will continue running at higher speed until it brings the heat sink temperature down, gracefully slowing down as things cool. U2 need not be mounted on the printed circuit board, but may be mounted remotely, on the heat sink; twist the wires together, and consider adding ferrite beads if there is a lot of RF floating around. Of course, controlling fan speed won't do much good if there isn't adequate cooling with the fan running at full speed. If you are adding a fan, size, placement, and airflow are important. For cooling a heat sink, impingement cooling, with the air blasting directly into the fins (like a Pentium cooling fan), is much more effective than ordinary convection cooling, where the airflow is just passing through the fins. If you are just cooling a cabinet or enclosure, sucking may be more effective than blowing. But any airflow is better than none at all.



Chapter 11

Propagation Topics

William Hepburn's VHF/UHF Tropospheric Ducting Forecast

Created By William R. Hepburn

William Hepburn has kindly allowed us to reproduce the following from his Web Site
http://www.iprimus.ca/~hepburnw/tropo_nwe.html
The Hepburn Tropo Index (HTI)
The HTI is the degree of tropospheric bending forecast to occur over a particular area, which is an indication of the overall strength of Tropospheric DX conditions on a linear scale from 0 to 10."

Negative Tropo Index = Below normal conditions. Bending occurs, but skyward.

Tropo Index of 0 = Normal midday "dead-band" conditions (Standard Atmosphere). Tropo Scatter only.

Tropo Index of 1 = Some downward bending occurs, but usually no discernible tropo.

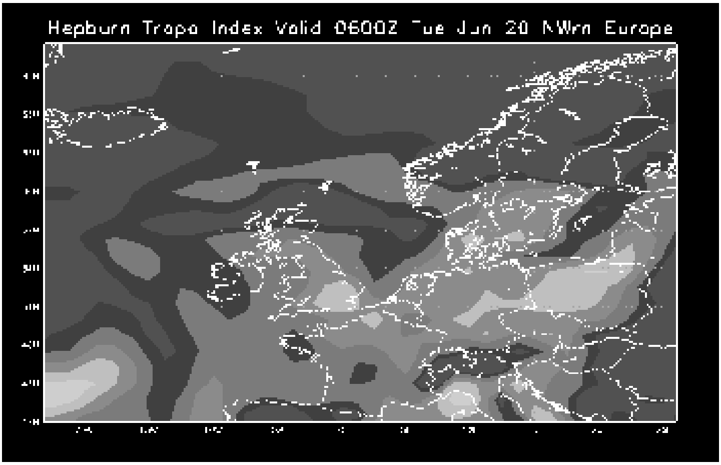
Tropo Index of 1.4 = Seems to be the average threshold of discernible tropo.

Tropo Index of 2 = Weak opening.
Tropo Index of 3 = Fair opening.
Tropo Index of 4 or 5 = Moderate opening.

Tropo Index of 6 or 7 = Strong opening.

Tropo Index of 8 = Very strong opening.

Tropo Index of 9 or 10 = Extremely strong opening.
(Yes, the Tropo Index can indeed exceed 10!...it is not a finite scale).



Background

I first attempted to receive distant TV signals in 1970 as a kid playing around with rabbit ears in my hometown in Southern Ontario. My first DX (tropo) was seen in 1972 when I received Channel 11 Toledo before our local Channel 11 came on the air. I started keeping a log of stations received on July 25, 1976. I had no knowledge of DX modes or that there were others who DX'd. On that 1st morning, I broke the Toledo record by picking up Channel 6 Columbus. Then on my 3rd day of organised DXing, I received Channel 3 Pensacola, FL (to my dad's amazement), destroying the Columbus record. (I wrote to the address given in an news editorial and was pleasantly surprised to find a WEAR-TV QSL Card in the mail the next week!). After picking up Florida from Canada, I wondered just what the limit was (of course, that was E-skip).

I later learned more about DXing from a copy of Communications World" magazine. In that magazine, I learned of the "World-wide TV-FM DX Association" (which I joined in early 1977) and learned about the various modes and what others were picking up. In 1981, I got my first taste of F2 skip with TV audio from the UK & Franc, and probable video from Ireland.

Other modes such as Meteor Scatter and Aurora were stumbled upon as the years passed. Forecasting Tropo. Using my many years of experience as a professional meteorologist with Environment Canada and an avid TV/radio DX enthusiast...I began experimental text DX forecasts for the Toronto-Buffalo area in 1997 (using E-mail). Once the Internet & newsgroups caught on, the forecast coverage eventually grew to cover North America. With continual gains in experience and real-time feedback from DXers, I was able to refine my methods and to gradually improve the quality of the forecasts. Later I developed the Hepburn Tropo Index to attempt to quantify the strength of Tropo Ducting Areas. Over time, the Index has been adjusted and refined and now does a decent job of representing the potential strength of Ducting. Then in May 2000, I prepared my first forecast maps. I automated the forecasts by having a computer program emulate what I had been doing manually to prepare the text forecasts and now have the computer prepare the maps. As a result of this automation, I have been able to expand the forecasts' coverage to encompass much of the world.

The Tropo Index alone is not an indicator of whether or not there will DX. The HTI is only indicative of favourable conditions overhead. In order to receive distant DX, the HTI will need to be relatively high over a large area. The larger the area, the longer the potential DX paths. Ducting paths will be straight-line (or Great Circle) only. Remember also that the longer the path, the more signal attenuation becomes a factor. Therefore for very long paths, a higher Index will be needed than that required for shorter paths. If an area's HTI is above normal, it indicates that conditions are favourable in that area.

The actual combination of weather conditions on surface and aloft will be the final factor in determining whether or not DX will be received in any particular locality. As well, physical barriers such as large hills or mountains may make reception via ducting practically impossible in some directions.

The Tropo Index does not factor in variations in Tropospheric Scatter, only Tropospheric Bending (Enhancement & Ducting). In the absence of any enhancement, and with only ever-present weak Scatter, the Tropo Index will be near 0. Also, don't be too concerned with the exact value of the Tropo

Index at your exact location. The purpose of the Index is to show potential duct paths (it is NOT a reading of the "probability" of ducting). Often, if you are within line-of-sight or within tropo scatter range of a duct located nearby, you may get in on that duct.

Tropospheric DX Modes: "Tropo"
The modes are defined by the mechanics behind them. A tropo DX mode is any condition that scatters, reflects or refracts signals in the Troposphere allowing DX to occur. Refraction occurs when the normal Index of Refraction has been altered. Vertical boundaries between different types of air masses usually cause this, where a temperature inversion (warm air over cooler air) exists. However, the most important influencing factor is water vapour (humidity). Thus, a warm dry air mass on top of a cooler humid air mass produces the best conditions. Dry Mexican air flowing across the Gulf of Mexico or Dry Saharan air flowing across the Mediterranean are two examples of prime tropo-producing conditions.

High pressure subsidence (the sinking and drying out of air), if it occurs over the oceans, can produce reception across several thousands of km! Hawaii to California reception, both on UHF and VHF, is not as uncommon as one might think. On the other hand, high mountains can physically block tropo DX, and deserts are generally too dry for tropo. Thus, tropo is rare in the very mountainous or dry regions of the world.

As far as classifying DX, enhancement and ducting in particular form a grey line. As a rule of thumb, enhancement is DX via inversions below 450 m (1500 ft) above ground...Ducting is DX via inversions above 450 m. (The layer of the troposphere below 450 m is called the "boundary layer" in meteorology).

Tropospheric V/UHF DX Modes.. Line-of-Sight (GW)
is normal continuous reception where the receiving and transmitting antennas can see each other, taking into account the 4/3 Earth curvature of radio waves.

Tropospheric Scatter (TrS)
is ever-present under normal conditions. That's the mode that produces the distant fluttery signals that randomly fade in and out. These are your most distant regular stations that barely make it in.

Depending on your location and equipment, tropo scatter can extend to 300..500..or even 700 km. The theoretical maximum limit for most TV/radio DXers is 800 km (500 miles) (Some semi-professional set-ups can extend further). Scatter is caused by small particles/droplets in the air such as haze, dust, volcanic ash, clouds, etc.

Tropospheric Enhancement (TrE), (aka Tropospheric Refraction) is common under normal conditions. On most clear nights, the ground radiates and the air near the ground cools. Eventually an inversion is formed and signals begin to refract off the inversion. Stations that normally fade in and out via tropo scatter come in continuously, with increasing strength. Also, weaker tropo scatter stations that are normally not heard (because their signal strengths never cross the background noise threshold signal level) also begin to appear. When the sun comes up, the ground & air heats up, the inversion breaks down, and the enhancement disappears. The enhancement is subtle on some nights, and very obvious on other nights. Distances are no different than your tropo scatter catches, it's just that the signals are stronger. Tropo enhancement is greatly influenced by terrain, with valley and coastal paths favoured.

("Fog-prone" areas are also "DX-prone" areas!!). From a DXers
point of view, multiple directions usually are enhanced at the same time.

Tropospheric Ducting (TrD)
is an abnormal condition. An inversion has formed at a much higher level above the ground...the vast majority of duct-producing inversions lie between 450 and 1500 m (1500 to 5000 ft) with a few between 1500 and 3000 m (5000 to 10,000 ft). These inversions are not formed due to night time radiation/cooling, but rather because of some other weather phenomenon (high pressure subsidence aloft, warm frontal boundary, cold frontal boundary, oceanic or lake inversion, Chinooks, etc.). Because of this, ducting can occur day or night (though it strengthens at night), is not usually influenced by terrain (East of the Rockies), and from a DXers point of view is usually either uni- or bidirectional. In fact, typical ducts are sharply directional. Signals refract off of and also travel along the inversion, thus the analogy of a duct. Distances are theoretically unlimited. One large area can have multiple ducts going on simultaneously, but they are usually parallel paths. It is possible in a very strong high pressure system to have large areas of ducting creating multi-directional openings. These are the rare "blockbuster" openings that make DXers' mouths water.

Additional Characteristics of Ducting. Ducting may or may not occur simultaneously with enhancement (caused by night-time cooling). Often there is both a low-level radiational inversion caused by night-time cooling producing enhancement)...and a mid-level "system-produced" inversion above that (producing ducting). However, just as often there is only the higher duct-producing inversion, especially if the skies are cloudy or if it is windy. So, do not use your regular scatter/enhancement stations as propagation beacons for longer distance DX achieved via ducting! Sometimes ducting can even display a "skip-like" character where distant stations on the same frequency and bearing can be received while closer-by stations are nowhere to be seen.

Ducting is also very height selective, with maximum signal transmissions at and just below the altitude of the inversion. Side lobes (what most ground-based DXers see from ducts) are similarly directional and narrow. Thus, conditions usually vary over short time periods as opposed to enhancement which is more stable. Ducts located behind cold fronts ("post-frontal ducts") are notoriously unstable as paths can even be interrupted by things such as heavy rain showers associated with the cold front itself.

Expect the unexpected from these types of ducts with sudden and rapid changes in signal strengths quite common (some post-frontal ducts last only 15 to 30 minutes). High-pressure and oceanic ducts are a bit more stable and can last for days, but again expect the unexpected as changes can occur quickly. Frequencies affected by ducting are determined by the vertical thickness of an inversion. Individual ducts will have a LUF (Lowest Usable Frequency) associated with them. Thin inversions (i.e.-thin ducts) will only propagate Microwaves. Thicker inversions will propagate UHF signals as well, while the thickest inversions will also propagate VHF signals. Unfortunately there is no reliable method known for forecasting inversion thickness’. See LUF page. Special Cases (Exotic DX Modes)..

Rain Scatter (RS)
is a rare mode that sometimes occurs on the higher UHF-TV channels. A band of very heavy rain (or rain and hail) at a distance can scatter or even reflect signals. The effect is the one used for microwave Weather Radar. Distances

are typically around 160 km, though up to 650 km (400 miles) is theoretically possible. (Note that heavy snow is not an useful reflector).

Ice Pellet Scatter (SS)..(called Sleet Scatter in the US)
is similar to Rain Scatter but is caused by bands of Ice Pellets in the wintertime.

Aircraft Scatter (AS)..(aka Tropospheric Reflection)
is simply reflection off of aircraft, although reflections off of flocks of birds are also possible. A rare form of reflection is "Chaff Scatter". Chaff is strips of metal foil sent out by the military during training exercises. Chaff helps to confuse enemy radar, but also helps to produce DX. Maximum distances for all reflection modes are again up to 800 km (500 miles).

Lightning Scatter (LS)
is a mode that is sometimes discussed, but there is little documentation on it. The theory is that lightning strikes produce ionised trails. Reception is similar to other forms of scatter except that the DX is more burst-like similar to MS. LS is a mode that is very hard to distinguish and rarely reported. Reflections off of hills and mountains, and Knife-Edge Diffraction are not considered true DX modes since they are Omni-present, though they can help to extend DX via the other modes.

So these are the conditions in the troposphere that allow reception of VHF and UHF signals beyond their normal range. Basically, these are DX modes that are affected by the weather.

In the forecasts, the reason that I stick with just Ducting is because it is a large-scale phenomena, can be put down in a forecast in a reasonable amount of time, and produces the best tropo. Enhancement is forecastable, but it is so dependant on regional and local terrain and conditions that it would be a labour-intensive effort (not to mention a very lengthy one) to forecast for all of North America. The process that one would have to use to properly forecast Enhancement for a particular area is the same that a meteorologist would use to forecast overnight low temperatures, chance of fog patches, etc.

There are also conditions in the ionosphere that produce distant reception via a whole different set of modes. Ionospheric "skip" and scatter are not caused by the weather, but instead by the interaction between the Sun and the Earth's outer atmosphere, or by objects such as meteors. For information on these modes, consult the ARRL Handbook.

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A 'New' Propagation Mode

Some observations by Phil Scorah, GW3PPF

Editor's comment:
If you will pardon the pun, a lot of water has flowed under the bridge since this article was written and the existence of sea ducts are pretty well established . I thought I'd include it though, as well as Martyn Vincent, G3UKV's reply. for historical interest.

With great reluctance, I have to concede the existence of a propagation mode that lies somewhere between Super Refraction (SR) and Tropo Ducting (TD).

For a few years now, Simon, G3LQR, has been reporting what he calls "Water Ducts" (WD), across the North Sea to Holland, that last all day and night. Penetration inland is limited but is significantly further than SR appears to go. Martyn, G3UKV has noticed that, during the Telford Microwave Expeditions, when good propagation occurs over the sea, it lasts all day. I have tried to explain it all away as TD with the occasional bit of SR but justification has become increasingly difficult with time!

Super Refraction has been hard to study since the move to narrowband operation on most of the microwave bands. All the suitably sited beacons are still on wideband! The arrival of Eric's' (F1GHB) beacons in Brittany has afforded a fresh opportunity to take SR measurements. Although the beacons are well inland, I suspect that they are RF line of sight to the sea beyond the North Coast of Brittany.

On the 12th July 1999, Roy, G3FYX, and I went to Seaton, Cornwall, IO70TI, (not the one in Devon). Whilst I spent all day measuring the F1XA0 and F1XAP beacons, Roy listened for F1XAQ mostly but failed to hear it. The graphs of our results (figs.1 and 2) show the standard SR shape, with the classic early evening peak.

On the morning of the 4th of September 1999, I was expecting Tropo Ducting and drove to my new site of Cefn y-Galchen IO81LSl9 (Mynydd Maen is no longer available to me). Finding no TD, I made a spur of the moment decision to return to Seaton and hope for a better SR situation than I found on 12.7.99, It was obvious from "switch on" that this was a much superior duct and I was hoping that the early evening peak would approach the Perfect Duct value. In the event, there was no evening peak and the awful truth slowly dawned on me ... that I had found one of Simons' infamous water ducts! Running in parallel with the SR and WD observations of XAO and XAP, I was monitoring the Guernsey 70cm Repeater GB3GU (vertical polarisation). It was not heard at all during the SR session but was heard, albeit weakly, during the evening only of the WD session. SR is renowned for only occasionally propagating 70cm but it usually propagates 23cm signals. It looks as though Water Ducts will be similar. More observations are needed, especially from those involving accurate measurements. I have some unpublished SR data from around 25 years ago and I am currently dredging through it trying to identify and WDs. Although Simon, G3LQR, is somewhat reticent on the subject of Water Ducts, the credit for their discovery is undoubtedly his, and quite rightly so!

Congratulations Simon!

Commentary on GW3PPF's article

Martyn Vincent, G3UKV

For as long as I've been active on 10GHz, 20 years plus, super-diffraction (or refraction, however you wish to name it) across the sea has been a recognised mode of propagation. The gurus of propagation led me to believe that stations at each end of a link had to be **in** the duct, i.e. on the beach or low cliff at each end, to use this mode. However, every time we've been portable at a sea location, we've found enhanced signals at 10GHz, notably from G3FNQ's personal beacon on 10368.2 MHz. Typically, signals start building up in the late afternoon around 1700 hrs, and peak soon after dark - typically 2200 in the summer. They then remain so until we go to bed! In the morning, they usually have returned to normal. Some days, no enhancement takes place, but on other days it remains all day. Generally settled weather gives rise to best SD propagation, as you would expect. Whilst portable on the cliffs in Guernsey in 1994, I was urged to go down from the cliffs (about 200ft high) to take advantage of SD, so we took all the gear apart and moved base about 1/4 mile to be on the beach: result ? Identical signal strengths from the GB3SCX beacon which showed the typical diurnal changes of SD on Guernsey.

So, for my money, there is only one mode involved giving enhancement on sea paths. It relies on stable temperature differences between the sea and the air above it. My results rely on evidence accrued in summer weather conditions, and may not apply at cooler times of the year. So long as there is a visual path to the sea, the duct can be used from any altitude. The name we give this propagation is irrelevant - but as super diffraction came first as far as I am aware, I'd settle for that as a label. Its enhancement properties are far more frequent and reliable than overland non line-of-sight paths. Personally, I'd settle for a QTH on the summit of a 1000 metre ASL hill by the sea, anytime !!

47GHz Propagation Observations

John Hazell, G8ACE

Propagation at 47GHZ is interesting to the writer insofar that, even after doing all the path profile work, signal exchanges can be equally poor or remarkable! Monitoring signal levels between particular locations is often only for a few hours, then not again for perhaps some months or longer.

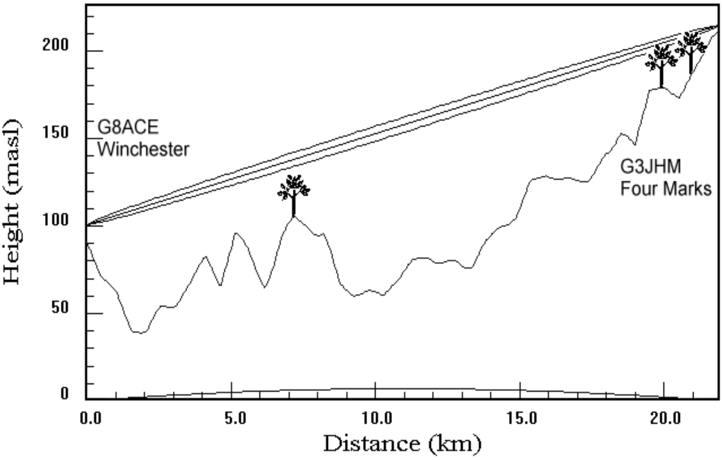
A beacon could be a solution to studying propagation variations, if one existed that could be heard at the home QTH of a 47GHz equipped microwave station. Looking at the profiles from local beacon sites to equipped local stations quickly revealed this as much of a non-starter.

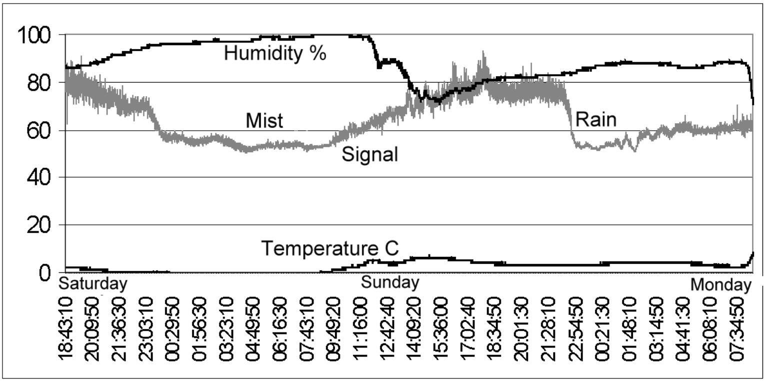
G3JHM at Four Marks is 22km distant from G8ACE in Winchester. Most of the path from G8ACE is good, as it sets off over the Itchen Valley, the Four Marks area being visible on most days. A QSO on 24GHz has already been conducted from dishes on tripods, garden to garden, so it looked optimistic for tests. G3JHM, having interests in propagation, was also keen to see if a 47GHz link could be established successfully.

A personal beacon was constructed and installed on the chimney at G8ACE in Winchester. At Medstead, (home of GB2RS in the south) just to the north of Four Marks, an excellent signal was receivable on the transverter, without any aerial, the unit being hand held. This was a true LOS path from Winchester, G8ACE‘s nearby landmark/ microwave reflector (HM Prison !) being clearly visible.

Moving south to G3JHM, the results were quite disappointing. The signal could just be found using a 45cm dish. The signal was spread over an angle of around 10 degrees with multiple peaks dependant on the elevation used. This is due to the large amount of vegetation clutter from trees and bushes in the last kilometre of the

The path profile showing the signal obstruction caused by vegetation clutter at G3JHM:





Signal level variation over a 36 hour period

path. The small horn initially used in the transmitter was replaced with a 30cm dish and the receiver for reception at G3JHM was constructed using a 60cm dish. This improved the signal so that good quieting can now be obtained using FM in "normal weather" conditions.

The RF Equipment:
The modules necessary for the transmitter and receiver were either specially built or already in the "cumhandy" box. The transmitter uses a G8ACE MKII ovened oscillator so that the frequency can be easily dc controlled with a multi turn pot. The FSK ident signal is added to this control voltage. My thanks go to Chris, G8BKE, for the building and programming the PIC ident generator. This drives a modified G4DDK multiplier to provide in excess of 20mw at 2.4GHz. With the enhanced 2.4GHz drive the multiplier to 11.7GHz uses just two GaAsFETs. The G3WDG009 is modified by removing the input mmic and replacing with a 3db pad. The multiplier GaAsFET is used as in the original design and the first amplifier is used alone to provide 10mW output at 11.7GHz. This simplified 2 GaAsFET version requires 7mW of 2.4GHz drive for full 10mW output using a 5v rail and recovered "Blue Cap" GaAsFETs. 10mW is more than adequate at this frequency for driving further multipliers. The 47GHz signal is achieved from three NE325 GaAsFETs. A doubler to 23.5GHz, then an amplifier is followed by further doubler. Between 5 and 10mW output can be achieved from this module for 2-3mW drive at 11.7GHz. Since the drive requirement is modest, the module is mounted behind the dish to minimise 47GHz losses and fed via 2m of 0.141 rigid coax at 11.7GHz from the original beacon unit now in the loft.

The receiver uses the same module arrangement as the transmitter to reach 11.7GHz. The 47GHz receive mixer is the original transverter built for the band by G8ACE and is similar to the DB6NT design.
Now a spare, it is utilised to save construction time. All the modules are mounted in a weatherproof box behind the receive dish with the signal leaving at 145MHz, the first IF frequency.
The IF uses a simple FM strip containing a 10.7Mhz crystal filter. The 2nd local oscillator is fixed tuned with a

crystal. Tuning of the 47GHz signal is achieved by DC control to the outdoor unit (97MHz) 1st ovened oscillator. The RSSI output from the FM demodulator is used to drive the data logging equipment.

Dishes (see photos on the next page)
Weather proof dish feeds needed to be created for this project and the arrangement is the result of some experimentation. It was found that 40mm Marley plastic waste pipe, slid over a splash plate feed, had negligible effect on the gain of the dish. The feed is constructed using 6mm o.d. copper tubing with a bore size similar to the usual 4mm circular waveguide used on 47GHz. This is fitted through a boss in the centre of the dish. The reflector is mounted at the far end of the plastic tube and the tube adjusted back and forth on the centre boss for focus. The tube is sealed with silicone sealant at either end. The wave guide then passes directly into the equipment box on the back of the dish.

Data logging:
A trawl of the Internet did not reveal any suitable software for signal logging purposes. Also the ability of G8ACE at writing PC software is non-existent. Equally, an A-D chip was not to be found in the cum-handy box. The BBC micro, long forgotten by most of us now, is easy to program and contains a four channel A-D. I should mention at this point a diversion into using the games port on a PC resulted only in wasted time as a practical test revealed each channel resolves only 16 levels … fine for a joystick but no use here. The BBC was duly programmed to draw a graph of received signal strength and record the levels to file. By chance, a screen dump program was found, thus allowing the graph to be printed. The dot matrix printer used, however lacked the quality we have become accustomed to and the problem of how to extract the data from the BBC still existed. A further Web trawl only produced complex answers to this problem; maybe I use the wrong search criteria?

Transferring serial data to the PC is easy, I finally discovered. Windows Hyper Terminal will log incoming Com Port data to a log file. Thus few lines of programme to the BBC and the data was leaving the BBC RS423 port to be easily transferred into Excel for a higher quality graphical display, analysis and print. A humidity detector and thermometer have been added to the data logging and three channels of data are now presented to the PC for saving to hard disk. A DOS terminal program was discovered so that data logging currently consists of the BBC for acquisition, sample interval selection and graphical display of signal in real time and an old 286PC for storing the files. G3JHM has undertaken to analyse the data and will produce the long term conclusions about propagation on the path. Already it is possible to readily see the effects of rain and mist. Some deep fades being noted. The project has entailed about 4-5 months of precious construction time but is enabling some propagation monitoring, perhaps not done before on this amateur band.

60cm Rx dish used at G3JHM

The 30cm Tx dish

A Digital Weak Signal Mode for Microwaves?

Chris Bartram, GW4DGU

Returning to an active interest in microwaves over the last year or so, I've been struck by the way in which there has been a drift of formerly active microwave people to VLF. While I appreciate the challenges of longwave, and could even share the enthusiasm of a friend, professionally involved in RF power amplifier design, who points out that he can legally and easily build 10+kW power amplifiers for 136kHz(!), I feel that the bands above 1GHz still offer some serious challenges. This doesn't necessarily mean translating existing (dated!) technologies to higher and higher frequencies, interesting as this can be.

An area where the VLFers have made progress has been in the application of very narrow band modulation schemes, allowing them to combat the inherent high antenna noise temperatures and small antenna efficiencies at those frequencies. It's easy to do this at VLF where the transmission path has relative phase and amplitude stability.

It's still possible to do similar things at VHF: Joe Taylor, W1JT, has demonstrated that his JT44 multi-tone FSK modulation (anyone remember Piccolo?) scheme based on the work of Bob Larkin W7PUA, operating in 5.8Hz demodulation bandwidths, has been effective for 50 and 144MHz and, to a lesser extent, 432MHz and 1296MHz EME using relatively small antennas. Bandwidth reduction also has applications on the microwave bands but there's a snag: microwave transmission paths, particularly beyond line-of sight, tend to be anything other than phase and amplitude stable! Poor phase and amplitude stability results in the spectrum of the signal spreading: quantifying this isn't a precise art, but tropo scatter at 1.3GHz results in a received bandwidth of perhaps 30Hz, whilst 24GHz EME has been reported to show spreads of maybe 250Hz. Conventionally this limits the minimum usable bandwidth to those figures: reducing the bandwidth further will decrease the signal to-noise ratio, or conversely, increase the receiver threshold! Human ear/brain processing can sometimes recover data from decorrelated signals but even that has limits. If a way can be found of reducing the decorrelation, the benefits could be very significant. 13dB system gain could potentially be obtained by reducing the decorrelation bandwidth of a 10GHz EME signal from, say, 100Hz to 5Hz. Potentially, this could allow contacts between stations using good quality tropo systems (5 - 10W and 100K receivers with cheap(ish!) 1.5m offset dishes using off-the-shelf satellite TV positioners, making 10GHz EME a very attractive proposition to a large number of operators.

So how can we do that? The key seems to be the use of a modulation format which includes a reference vector. A reference vector is simply a component of the modulating signal of which the receiving system has a priori knowledge. It could be as simple as a constant carrier. If the received reference vector can be reconstructed, then as we are considering a narrowband signal, the complex function required to reconstruct the reference can be applied to the whole signal spectrum, and data can then be recovered in narrow bandwidths. There's nothing really new in this. A similar technique has been very successfully used in the Linear Modulation (LM) systems developed for ssb speech and data transmission in mobile radio environments. Incidentally, the use of a reference vector isn't the same as the use of a reference frequency in the JT44 modulation format. That provides a non-continuous frequency reference: I'm suggesting that the reference vector is transmitted continuously to provide a coherent reference.

Although some elegant solutions, such as various forms of PSK, and even QAM, exist, multi-level FSK combined with a reference vector could provide a simple solution. This wouldn't give a constant amplitude signal - which may or may not be desirable - but it would work with standard ssb transmitters, albeit probably generating some transmit intermodulation. It presents a similar challenge in terms of transmitter linearity to PSK31. Whether this is significant (most of the time!) on the microwave bands is a moot point! There are a number transmitter linearisation techniques, particularly Cartesian feedback loop topologies, available if it's necessary to clean-up the intermods. These ideas need testing by proper modelling (Matlab, anyone?) and maybe then by building a test system based around PC sound cards. I can't see myself having the time to do this in the next few months as I'm involved in a protracted house-move and I'm also pretty busy in my work. I'd also be on a steep learning curve in certain areas. I'd be very happy to co-operate with anyone with the necessary skills. I believe algorithms exist to allow reconstruction of the reference vector and the other signal processing tasks are if not trivial, not the DSP equivalent of rocket science.

As the VLFers and K1JT/W7PUA have demonstrated, the use of PC based techniques to increase the capability of existing VLF to VHF systems is entirely practicable. Currently available software is of limited use in the microwave bands, and a properly thought-through system taking into account the properties of the microwave transmission medium is needed.

Rain scatter – Where, When, How?

Uffe Lindhardt, PA5DD

This paper formed the basis of a talk on rain scatter given at the Adastral Park Microwave Roundtable on 14 November 1999. It deals with the practical aspects of using rain clouds as a mean of reflection on frequencies around 10 GHz. It also includes a look back on the rain scatter season 1999 as seen from the Netherlands.

Frequencies

For radio amateurs the main frequency band of interest for rain scatter contacts is 10 GHz. Rain clouds offer very good reflectivity at this frequency, and atmospheric losses are low. At the same time the availability of high performance components and circuits for this band has increased the last few years.

On lower bands the reflectivity decreases drastically due to the size of raindrops compared to the wavelength. According to WA1MBA (see Internet resources below) this amounts to –12dB at 5.7 GHz and –19dB at 3.4 GHz in relation to 10GHz. These figures seem to match the practical experience of many amateurs. DX-contacts (over 400 km) are quite difficult on 5.7 GHz. Also the number of stations QRV on this band is much smaller.

Rain scatter on higher bands than 10 GHz is an area where very little experience has been gathered. Due to the still very merger activity and the small output powers on 24 GHz, very few contacts has been made. Certainly atmospheric losses start to play a role on this band. It seems though, that shorter contacts of up to 200 km can be made when strong forward scatter is present on 10 GHz. For these contacts elevation is essential.

Equipment

Any equipment used to make tropospheric contacts will do, but to make full advantage of this propagation mode a narrowband mast mounted home station is essential. The weather patterns that rain scatter are linked to, do not invite for hill top portable operation. It is also very difficult to predict when good rain scatter openings will occur, and it is therefore essential, to monitor the conditions over longer periods.

QTH

It is not essential to have a high-elevated QTH for working rain scatter, as it is often the case with tropospheric contacts. High altitude will give only little enhancement in the achievable ODXs. What is important is a clear horizon because even a few degrees of horizon elevation are going to block for DX contacts. The rain clouds reside only up to a maximum height of 10 – 12 km, and it is essential to have a clear view of the reflection areas on the horizon if you want to achieve contacts of 600 – 800 km.

TX

A reasonable ERP output power is important. Output powers of up to 1W are available at low cost (e.g. QUALCOMM surplus modules), and with that power and a parabolic dish of 50cm, you have a good rain scatter station. In fact most of the QSOs reported at the end of this paper were worked with 1W and a 45cm dish, including several QSOs of 600 – 700 km.

Since the station should preferably be mast mounted - because of the high feed losses at 10 GHz - solid-state amplifiers are easier to install.

On the other hand the cost of solid-state at a power outputs above 1W are still very expensive.

Installing a TWT amplifier in your mast can be quite cumbersome. Nevertheless many stations using 10W or more are QRV via rain scatter, so be prepared for some frustration if you go on the air with 200mW !

RX

Since low cost HEMTs like the NE325 are now available, the receiver should have a noise figure of 1 – 2dB. On the other hand since many high power stations are QRV any receiver will do.

Antenna

For maximum size/gain performance a parabolic dish of at least 40cm should be used. Using a relative small dish gives some advantages in finding the optimum reflection point and also removes the need to elevate the antenna at medium distance contacts.

A larger antenna gives the advantage of a larger ERP output power, which can prove essential for DX contacts. Exact pointing gets difficult from a dish size of 70cms onwards using the normal commercially available azimuth rotators. My advice is to use a 40 - 50cm dish if you are looking for many QSOs (like in a contest), and a larger (70 – 90cm) if you are a DXer.

Elevation can be very useful in rain scatter openings. For medium distances (300-500 km) it can mean a difference of 10 – 20 dB in signals. It is however my experience, that you will work these stations also without elevation but probably AFTER that the stations with elevation are done rag chewing. For 24GHz, elevation is essential for rain scatter.

Modes

All though rain scatter is possible using broadband equipment real DX requires the better system performance of narrowband modes like CW, SSB & narrowband FM. CW (telegraphy) is by far the most efficient mode. Due to the fast random orientation wind speeds in rain clouds, reflected signals are subject to Doppler distortion, which makes the signal sound like white noise. The size of the Doppler effect is on 10 GHz very similar to the one known from 144 MHz Aurora reflection. This distortion makes it difficult to understand SSB signals, especially when using side scatter (i.e. the reflection point being offset from the direct line between the two stations). On the other hand SSB is normally quite useful for making DX QSOs, which are always forward scatter.

For local or medium distance contacts - where signals can be very strong – narrowband FM is also an option. In FM the Doppler distortion disappears all together, due to the low deviation of the Doppler effect as compared with the modulation. For making fast comfortable QSOs or for rag chewing FM is the perfect mode, it is however recommended to QSY from the narrowband part of the band to prevent disturbance to other stations.

Finding rain scatter

Finding rain scatter openings are quite a challenge because of the sparse distribution of stations and beacons, and because of the narrow beam angles associated with 10GHz antennas. The rain scatter season – at our latitudes - extends from approximately begin of May to the end of September, with a peak in the month of June. Certainly rain scatter contacts can be made outside this season, but they rarely produce any DX contacts.

At present most of the rain scatter QSOs made are made along the 50º latitude. From my experience rain scatter is less frequent at higher latitudes like 55º. It is however difficult to separate the effect of the generally low 10 GHz activity level there and possibilities for rain scatter. There seems however to exist a belt of high thunder activity going from the Biscay bay and North- East towards central Europe. In any case there will certainly be some variation in rain scatter activity from region to region.

At present the main warning channel for rain scatter openings is the Packet Cluster network. *(or nowadays the ON4KST Chat site. Ed.)* This network links most of Europe together for fast real-time spotting of rain scatter observation. The ideal rain scatter spot contains information of the QTH of the two stations, and the azimuth angle QTF of the reporting station. Unfortunately only the newer DX cluster software has a good support for reporting 10 GHz contacts, and this software is not yet installed throughout Europe (e.g. not in the UK). For example CLX provides excellent features like dedicated 10 GHz spotting and reflection point calculation. WW-converse is also available via the Packet network. It is a "chat mode", and allows real-time communications between rain scatter stations on channel 10368.

Unfortunately WW-converse is also not available everywhere, and normally less than 10 stations van be found here during openings.

The Packet cluster & WW-converse can also be reached via various "back doors" on the Internet. So can some of the professional weather radar's. These radars typically work at around 9 GHz, and they can therefore provide very useful information on current rain scatter conditions. The main problems in using these sources are the cost of real-time information, which is sold at commercial terms. The information that is publicly available is normally some hours old, and does for example not include elevation scans, which are essential for the evaluation of DX possibilities. A few exemptions are mentioned under the Internet resources below. Beacons are another essential tool in finding good rain scatter reflection points. It is however essential with high power beacons (1W or more), since the beacons are required to have an Omni-directional antenna pattern, and hence have relative low ERP. It is also important that beacons are placed between places, where actual activity can be expected. A good example of such a beacon is DB0JK in JO30LX, which serves the three main areas of activity at present, namely the Randstad NL (JO22), Ruhr D (JO31) & Rhein/Main D (JN49). Unfortunately the lesser-activated areas (like the UK at present) suffer from the mutual concentration of activity between these centres. One answer to solve this problem is the deployment of beacons to attract attention.

A final means of locating rain scatter reflection points is the use of another local station as "sounder". In my experience the backscatter signal of a station close to you gives the best indication of the beam angle towards the DX stations. Beacons which are often received via side scatter will give a beam angle slightly offset to the forward scatter beam angle. A very good teamwork can evolve between two close-by stations alternately giving CQ, while the other station are optimising his beam angle.

Combined with some kind of backbone communication, this can be a powerful tool to find the reflection in directions, where no beacons are present. It can however be difficult to determine the distance to the scatter point using this technique. Maybe in the future amateurs will be able to do their own "sounding", using fast RX/TX switching and precise time mapping of the return signal.

Finally it cannot be stressed enough, that activity brings on more activity in quite a surprising way. It is my estimate that at present, there are close to a doubling of rain scatter activity every year on the continent, at least if measured by the number of contacts made. For an indication of the present activity, you can check the list of stations compiled by DG1VL and referenced below under Internet resources. This list is only the top of the iceberg!

Operating

The rain scatter activity is concentrated around 10368.100 MHz. During openings the activity can extend well beyond the band 10368.080 – 10368.150 MHz. It is apparent,that more spectrum will be needed if activity continue to increase at the present level.

It is not unheard of to hear 10 stations giving CQ at the same time in this band segment during an opening. Most contacts are made randomly following a CQ without any prior arrangement. This makes rain scatter contacts an extra treat, since most other contacts on 10 GHz are made after contacts made on lower bands. The normal operation style is to have an automatic keyer make a CQ in the direction, where the rain scatter reflection point is assumed to be. These CQ calls can be quite long but ,in order to find the rain scatter, it is important that the on-air time is kept high. Unfortunately this leads some operators to call CQ for more than 10 min, without checking for stations coming back to their call. If this happen to be the rare DX station you are looking for, be sure to have some tranquillisers at hand! Calls in SSB are also made, but usually even stations that cannot read CW uses an automatic keyer to make the CQ.

A main activity during a rain scatter opening is to keep track of the reflection area. The reflection point is often moving (though not fast), and DX stations will have slight variations in the optimal azimuth angle, depending on the angle they have towards the reflection area. In some big openings there are 2 or 3 different reflection areas, which can be challenging to the operator. A good working Packet cluster can help to focus on the best direction.

At the IARU conference in Lillehammer 1999 it was decided to replace the last character of the RST report with an S (e.g. 59S) in rain scatter contacts. This reporting is used on CW as well as phone. Log analysis 1999 To round off this paper I have made some simple analysis on the rain scatter contacts that I have made during 1999. All of the contacts were made from my home QTH that is situated at –2m ASL, but with 360º of free horizon take-off. Most of the contacts were made with 1W output and a 45cm parabolic dish (actually a lampshade). Towards the end of the period I have upgraded to 10W output and a 70cm dish. These improvements were made as an effect of the good results made with the original set-up.

The first figure in appendix 1 shows the distribution of contacts and distances over the season. Of course the number of contacts are highly dependent on when I was actually QRV, but still the figure are rather interesting. It should be noted that many stations have been worked more than once. For a detailed analysis I refer to the complete log in appendix 2.

The first figure shows 5 – 6 real DX openings concentrated in the period May, June and July. The DX openings are characterised in that contacts over 450 km are possible. These contacts are much rarer than contacts of 200 – 400 km.

This is shown more clearly in the second figure, where the same contacts has been sorted after distance. DX contacts are more difficult to achieve due to the fact that openings are shorter, and DX openings can rarely be detected using beacons. The signal strengths of DX stations can be quite impressive though.

Conclusion

It is my hope that this paper shows that rain scatter reflection on 10 GHz is a mode that everybody can enjoy, and that we can continue the rising trend of rain scatter activity in Europe in the coming years. Today the achievable DX results on 10 GHz far succeed those of the lower microwave bands like 2.3 GHz. This is a surprising development, from which we have only seen the beginning yet.

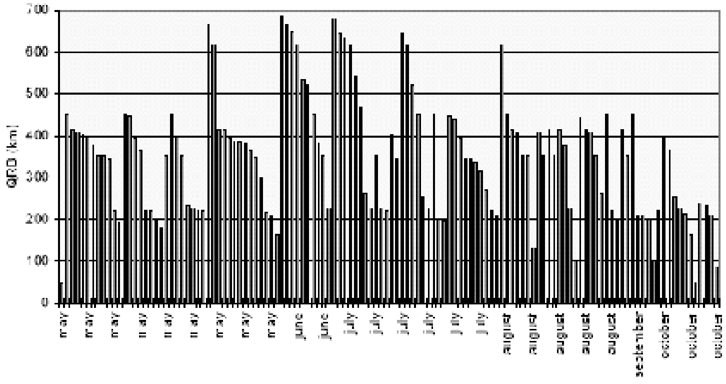
Internet resources WA1MBA, on the basics of rain scatter:
http://www.wa1mba.org/10grain.htm

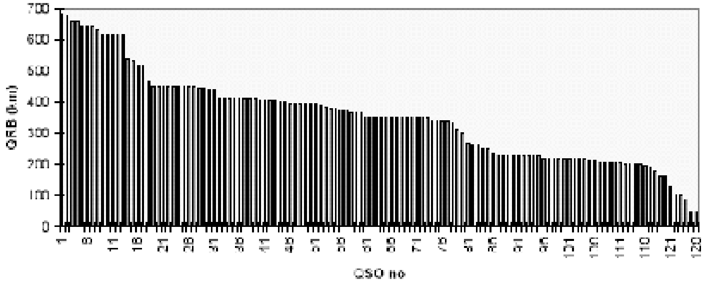
DG1VLs list of rain scatter stations:
http://www.qsl.net/dg1vl/RS_05_02_99.txt

Weather radar for the Netherlands:
http://weerkamer.nl/radar/
Weather radar for Bonn, Germany with elevation scans:
http://www.meteo.uni-bonn.de/Deutsch/Forschung/Gruppen/radar/radar_en.html

PA5DD, Authors homepage:
http://home.worldonline.nl/~nouchavw/

Appendix 1





5760 MHz Rain Scatter Detector

Rudi Wakolbinger OE5VRL

I have been making QSOs for many years using Rain Scatter (RS) on the microwave bands. Through tests with DL6NCI, DD7MH, HG1YA and many others, I have confirmed that 3cm is not the only band suitable for RS. It is also possible on 6cm (very good), 13cm (just about), and 24GHz (very rarely). 9cm is not available in Austria so my experience of that band is minimal.

To make RS contacts, you must first find a suitable cloud that can act as a scatter point. The finding of these clouds has until now been mainly through listening for beacons. With this method it is only possible to find the direction of the scatter point but not its distance, so it is not possible to determine the actual location of the scatter point. That brings me to my idea, to receive my own reflected signals and to measure the elapsed time. I have had this idea for years … it was recently re-activated through an article by LX1DU in DUBUS.

I must acknowledge my radio friends for solutions to several technical problems. Above all is my friend Erwin (OE5UXL) and I must not forget Michael (DB6NT) with his excellent modules and amplifiers. I had many QSOs with Ulrich (DG2MF) in which we discussed fundamentals. Ludwig (DC8NV) provided the most significant part, a circulator. Dieter (DL3NQ) gave me many important tips. Ferdi (DC8EC), Erhard (DC4RH), and many other helped with their specialised knowledge to help me realize this project. Based on my experience of RS and the availability of cheap components I decided to operate my Rain Scatter Detector (RSD) on 6cm, 5.7GHz.

I needed a number of units for the system. A pulse generator to produce the transmitted pulse and receiver phase, a transmitter oscillator (120MHz), a 48x multiplier to produce 5760MHz at 100mw, an 5W amplifier, a converter 5760MHz to 144MHz, a 144MHz receiver with a S-meter output and, finally, a circulator. An oscilloscope is used to display the results. The pulse generator produces a 30µs pulse with a frequency of 375Hz. This controls the transmitter that produces a 30µs carrier, 375 times a second. There is a single buffer amplifier after the oscillator and three multiplier stages in the transmitter driver. The transmitted signal is amplified by a two-stage amplifier up to the 5W level and through a circulator to the antenna (3m dish).

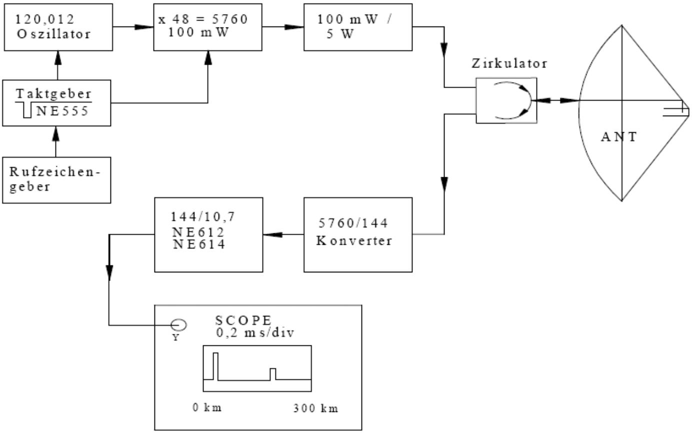
A very good match at the antenna is a basic requirement, otherwise too much reflected power may flow over the circulator to the receiver and result in the destruction of the input transistor. The isolation between the transmitter and receiver ports is 33dB. Approximately 3mW from my transmitter appears at the receiver input. During tests I determined that the input transistor (NE325) could survive this power level.

The antenna radiates the transmitted pulse and when no reflector is available, it disappears. A small amount of the transmitted signal is received via the circulator and this causes the oscilloscope to display a large Y deflection for the duration of the transmitted pulse. The transmission of the pulse triggers the start for the X time base. Suppose that within 300kms of my antenna is a good reflector, a proportion of the transmitted pulse is reflected and if strong enough, it will appear on the oscilloscope as a deflection on the Y axis at a point in time on the X axis.

The size of the Y deflection is proportional to the strength of the reflected signal and the distance to the reflector is calculated by measuring the time difference between the pulse being transmitted and the reflection being received. An elapsed time of 1ms = 300kms which is produced by 150km there + 150km return, indicating 150km to the scatter point. I have the X axis time base set to 0.2 ms/div so 300km are easily seen.

To date (beginning March), there have been no thunderstorms and that’s why I still do not have much experience of the range of my RSD.

The RSD first went into service in mid February and I easily received a reflection from a hill in the Swabian Mountains on an azimuth of 271 degrees and at a distance of approximately 370ks . In the north direction I can see a number of hills at around 200km distant. My RSD not only detects rain scatter, but also Over Rich Distance Detector ORDD. Over the summer months I hope to show how close to theory will be the results in practice. This system functions only when using a reasonably large antenna (3m dish). If I had only a 1m dish, then I would need a power of 500W to get the same strength from the reflections. With a 70cm dish I would need 1kW.



A Proposal for a Passive Rainscatter Monitoring System

Lehane Kellett, G8KMH

Introduction

The UK Microwave Group was recently discussing potential projects and the perennial discussion of beacons came up. In the UK, at least, it has become difficult to find acceptable sites or groups willing to take on the establishment of 10GHz beacons or for that matter beacons on other bands. There are already a number of beacons but these may not be in the best position to allow the determination of rain cells and hence a significant lost opportunity for 10GHz contacts.

Looking in reverse however, if there can't be enough beacons in the right locations then it might be possible to put receivers in sufficient locations to either receive a beacon or transmissions from stations probing to find scatter (or even tropo ducts).

Putting a receiver in place is one thing but there has to be some feedback path to the user(s). In an ideal world a nice high speed connection to the Internet would be available with FFT displays at the user end. In the real world the best we could achieve at present is an audio quality line from a GSM cellphone.

This paper discusses the approach for a simple beacon monitor and discusses what might be possible in the future. In essence the concept is to produce a self contained box which requires only power (240V AC or 12V DC) and siting on a mast or other convenient location.

Rain scatter monitoring

There is little point in covering what rain scatter is in this paper. There are plenty of references to it elsewhere. For the purposes of the monitoring system there are two key points

- The rain cells move fairly rapidly and therefore a path between two sites is only transient.

A system must therefore have a reasonably large beamwidth to cater for many path permutations.

- The signal suffers from dispersion due to the random Doppler from the raindrops.

Ideally a system would cover a wide bandwidth with digital signal processing of the signals.

Whilst we can provide for a reasonable beamwidth using a horn antenna (at 10GHz) – in fact their small size is a positive bonus!

However, it is not easy to provide a large bandwidth, say 25KHz, and send processed data down a 3KHz audio or 9600bps digital link. So, for the first phase the receiver will be no more than a standard SSB receiver with 2.4KHz bandwidth.

It may be possible to use a wider bandwidth and use some audio frequency folding - 4KHz becomes 0Hz, 6KHz translates to 2KHz, etc. This would best be done using DSP.

Phase 1 Remote receivers

Our first system is likely to use readily available components – a 15dB horn, LNA, 10GHz down converter and then a commercial (trans)receiver. It would be advantageous to create a modified receiver board (Plessey, Howes, etc) in place of the commercial unit, if only to reduce power requirements. The GSM phone is key to the operation of the system. It is not known if a similar system has been used commercially – probably given some of the roadside boxes with 900MHz antennas - but it is understood that the approach has been used in clandestine operations. It is not legally possible to take a cellphone, open it up, and attach wires to the keypad, reset lines, etc., although this would be easiest. Another approach has to be taken and some means of controlling the phone is required, as well as access to the normal hands-free / car-kit microphone and earphone connections.

Initially the search was for phones with a logic level ring indicator and answer system but this didn't prove fruitful. Next those with data connections were looked into and there are two main candidates – most Nokia and some Ericssons. The latter have an easy to use Hayes command set but crucially lack the ability to turn the phone on if it has been powered off (which may happen during a prolonged power outage). So a Nokia it is – probably a 6100 series. Unfortunately the protocol for the data connection is not published by Nokia (except under NDA) but has been reverse engineered by the Linux community and the details put on the Internet. The cellphone (or at least the SIM) will be one of the popular pre-pay systems with vouchers that are valid for a long period without the need to top up the phone. Some allow incoming calls for 6 months after the expiry of the voucher. A controller board will interface between the receiver and the GSM phone. It will provide for all the call control – answer, hang up after preset time, power on, etc. It will also have DTMF decoder for future options.

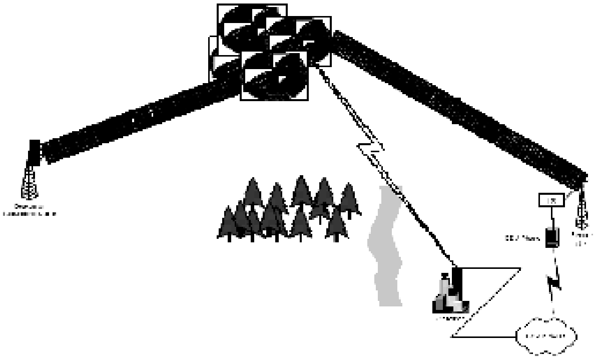
Phase 1 operation

In Phase 1 only the most basic operation is possible. Here the operator makes a call to the unit and then listens – either for other operators/ beacons or transmits his/herself. By using different beam headings it would be possible to 'probe' for areas of scatter. Of course, with sufficient units a local system may just be used to check you are still transmitting – very useful when there seems to be no-one else on the band!

Phase 1a,1b,2 – futures

Looking to the near future it should be easily possible to replace the commercial transceiver with a simple SSB receiver. It would then be possible to add some additional facilities, namely:

- Switchable LO crystals or PLL LO operation



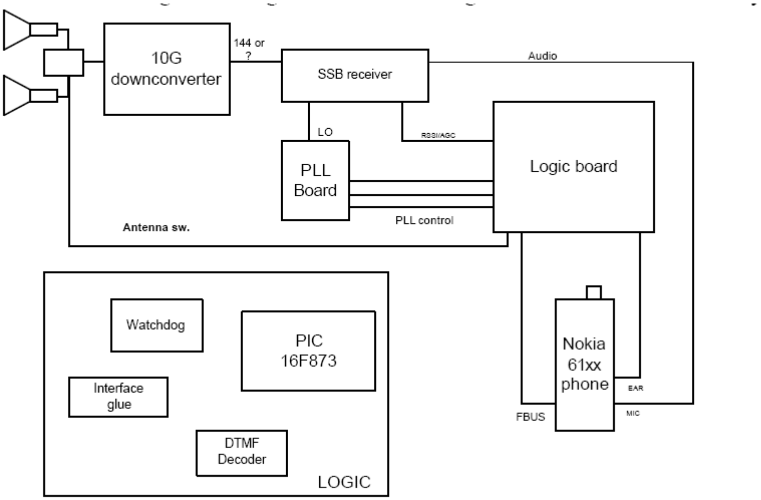
- Switchable filters
- RSSI/AGC A to D conversion
- Audio fold back (mixer and filter)

The PLL control of the LO would be very simple to add as most of the work has been done by Andy, G4JNT. All these functions would be controlled by means of DTMF commands. In the future it would be nice to add DSP capabilities and make the results available on the Internet via a push server. This may be possible if an academic establishment makes a site and bandwidth available.

With some care it should be possible to use a solar cell to provide the power, with the logic board monitoring the voltage levels and turning off the receiver if necessary.

Conclusion

A rain scattering system looks feasible and should be reproducible at low installed and running cost. How many could be established in the UK would depend on funding but certainly as many units as there are beacons should be a target. The use of a cellphone makes the unit almost completely self contained.



Chapter 12

Measurement

Testing Times

© Lehane Kellett G8KMH

Introduction

This is article covers a range of test equipment from simple diode detectors through to spectrum and network analysers, how to build them – diode detectors not network analysers! Where to buy, how much, etc. Most importantly I aim to cover their usage in a typical home workshop – not in some esoteric research lab.

Your first test lab

Let's admit it, part of the fun in microwaves is spending the time on the bench with a piece of equipment, trying to get that last extra bit of performance from it. I hate to think how many hours are spent getting half a dB more from amplifiers but to be able to do this you need a certain amount of test equipment. Building an LO chain for your 10G transverter needs a different set. Let's look at some of the things you'll find reallyuseful in say, building a 10G system. You probably have many of them already so you're half way there.

Tools

It may seem very basic but you will need some good quality hand tools. Don't buy junk which wears out or won't close properly. Personally I like Lindstrom but CK aren't too bad. Next you'll need a good soldering iron, preferably two. Try and get a temperature controlled one like the Weller TCP series. The second one can be a bigger one for soldering tinplate, etc. Get some 22swg solder and some SMD solder paste (RS/Electromail). A good 6" stainless steel rule comes in handy as does a magnifier. A good hand drill and a variety of twist drills are needed, from 0.4mm through to 10mm.

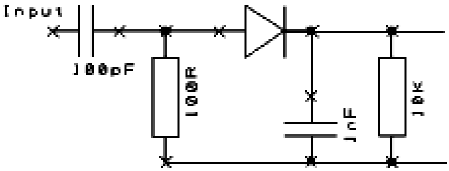
Meters

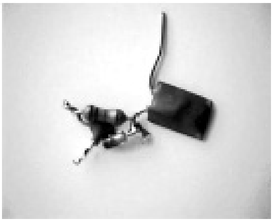
Seen those £9.99 DVMs you can get? Keep walking! Most of these aren't designed to give much in the way of accuracy and you really have to spend a little more to get a decent one (Having said that, a $4.99 one obtained in the States is to 0.1% - just luck). You don't need all the bells and whistles, just decent accuracy. Next you need an analogue meter, you just can't see the changes fast enough on a DVM to be able to align things. Old AVOs and other good meters go for £4 up at rallies.

RF

You may be thinking that you need stacks of expensive HP test gear to build your first 1.3/2.3GHz transverter or ATV transmitter? It can be done with two things – a homebrew RF detector and a homebrew wavemeter to 2.6GHz. I'll describe how to make them in a second. A few more can make things easier and I'd suggest a signal source, a surplus detector and some co-axial attenuators. I'm actually going to cheat and tell you how to make two detectors since you really can't describe the first one as requiring any effort.

To make a great RF probe, good enough to align your LO sources first off, you'll need a glass Schottky diode (HP HSCH2800 etc), two capacitors, one about 1-100pF (lower is better for microwave work) and the other 1nF, one 10K resistor, one 100R resistor and two bits of wires. The circuit below shows how it works but the photo tells the real story. Just twist all the bits together and cut off the leads, if you want you can use some solder! Costs about £2. If you are worried about





cosmetics you can package it in a pen, cigar tube, whatever! The second detector is more accurate and can be used up to several GHz. You'll need to make the PCB. This is basically the same circuit but using surface mount devices (SMD) – it'll be good practice since you'll be mounting lots of SMDs over time. Without all the stray reactance's from those leads it will have a fairly uniform response and the published graphs show a plot of power vs. output volts for different diode types.

The second item is a cavity wavemeter, very accurate and useful for checking the LO strip, typically a DDK004, is on the right frequency (typically within a few kHz). To build it you'll need to do some metalwork with either brass sheet (preferred) or tinplate. The device consists of a 3 sided long box with a 4 or 5mm brass/copper tube/rod in the middle. You can either mount it up on a piece of wood and calibrate it using a ruler or measure the depth the rod is in the cavity each time. Full constructional details of one based on a tube are in the RSGB Microwave Manual, Volume 2.

Of course, surplus commercial units are better and can be had for as little as £20.

The other optional items you can't make. That's the bad news, the good news is that commercial detectors and attenuators are about at rallies and round tables.

Commercial detectors typically come in SMA or N type connectors and cover a range up to 12.4 or 18GHz. Get attenuator(s) that match your detector, either N or SMA, and one or more adapters from N to SMA. If you can, get 3dB, 6dB and 10dB ones. Look for names like Weinschel, Narda, MidWest Microwave or, if you are lucky, Hewlett Packard (HP). A very useful one would be a 12.4GHz 10dB 2W. A good detector is from £5-15 and attenuators from £1-5 although the 10dB 2W one could be up to £25 – unless you are extremely careless it'll last a lifetime. Well, what haven't I mentioned? There are lots of things you can add to your 'lab' and I'll cover them later.

Scalar Network Analysers

Impressive sounding name! Many of you will have used something similar without really thinking about it. At its simplest it is an RF source which can be varied over a range of frequencies ('swept'), a detector with a logarithmic amplifier and an oscilloscope. Add a second detector and a directional coupler and you have a versatile tool.

Before you start thinking about how to build one, let me tell you that these units, without the RF source, come on to the surplus market for £25-60 without the detector head. In fact because they have no detector heads! Detectors, when available, tend to be about £75-300 each. Of course you may ask what use one is without the detectors? Well, mostly they can be used with homebrew detectors or with commercial N/SMA detectors and a few resistors or an op-amp. Except the HP network analysers which, in order to get a better dynamic range, use an AC coupled amplifier and AM modulate the RF source at 27.8KHz. Look for the Wiltron

560/560A and the Marconi 6500, unless you can get one with detectors in which case the HP8755 with an 182T is a nice unit. Later units, like the 8756/7, are probably outside most amateurs budgets at £800 up.

Typically the units require two interconnections between the sweep oscillator and the network analyser. One will be the sweep control voltage, a sawtooth/ramp waveform, from the oscillator to the analyser and this varies from 0 to –10V, 0 to –8V and +/-5V depending on model. Another is the blanking output which blanks the CRT display when the sweep ramp is decaying (retrace). On the HP units there will be the AM modulation input as well and some Wiltrons have a marker output. The source can be from a simple one transistor VCO up to a full blown HP sweep oscillator. HP and other sweepers are regularly available surplus and I'll cover sweepers later. Don't forget your old Gunn oscillator can be readily made into a sweep oscillator with the addition of an op-amp sawtooth generator and an op-amp level shifter. You can get by without blanking to start with, it looks messy but does the job. Having got one, what can you do with it? Well, all those filter alignments become easy with one of these – you can *see* the filter response. Think your 10G antenna isn't behaving? Sweep it with a Gunn and either a circulator or a directional coupler and see whether the reflected power is better higher or lower than 10368. Building a 23cm amplifier? Sweep it with a one transistor VCO and see how flat it is and whether it covers just 1296 or 1268 (satellite) as well.

Sweepers

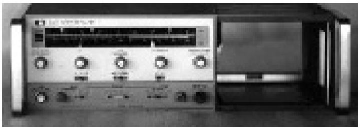
Some people may think this is my favourite topic! But if I had to keep one piece of professional RF equipment then this would be a sweeper. What's a sweeper? A signal generator where the frequency can be linearly varied over time by a ramp voltage. A super VCO. There's more to it than that, you can specify the start and end frequencies and the time taken for the sweep and some have markers so you can 'spot' a particular frequency, like the centre of a filter bandpass. As an aside, one of the professional magazines described the idealised filter response as the 'Barthead'! Think about it…

As I mentioned last time, sweepers are available if you keep your eyes open at rallies, from some of the dealers or on the Internet from the USA or Germany. In the UK probably the most common is the HP 8620 which is both quite small and very versatile since it accepts a range of HP oscillator plug-ins covering 10MHz to 22GHz. A basic 8620A without a plug in you'd pay between £30 and £60 for. A late model 'C with the options, such as HPIB, goes for up to £300. The bad news is that without one of the plug in oscillator modules it makes a reasonable door stop and not much else. The second piece of bad news is that there are quite a few non working units too! So choose carefully at rallies, that bargain may be useless or difficult to fix. The range of plug ins is listed on the website www.microwavers.org/sweepers for reference. There may be others and some options increased the range of the 18/22GHz units (usually options beginning H).

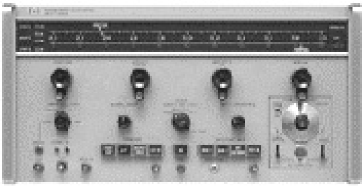
Another sweeper which you may find is the HP 8690. This is a larger unit than the 8620 and has a wider range of modules – even some 3rd party to 140GHz.

Many of the plug ins use Backward Wave Oscillators (BWOs) which have a finite life. So, even one working now may not have much left to go –see if it has an hours indicator on the chassis. The 8690 doesn't seem to appear as much as the 8620 in the UK but is more popular in the USA. However, shipping costs make it prohibitive to import and you could buy 2 8620's for the cost of the FedEx bill….but that's another story.

The 8620 was superseded by the 8350a/b. These are *really* nice digital units which accept an even larger range of plug ins than the 8620. Even better, with the 11869a adapter you can use your old 8620 oscillators in the new mainframe – just as well as the 835xx series are still fetching 4 figure sums after 15 years and you can buy a decent Skoda for one of the wide range 10MHz to 20GHz units. If you get the 11869a adapter then check if it has the loose BNC plug on the purple coax. This is the FM connection to the 8350 back panel and makes phase locking the unit possible without trailing coax through the back of the 11869a adapter. If it is missing you can get the coax connector for the D-type (Arrow?) and make your own with a length of RG174.



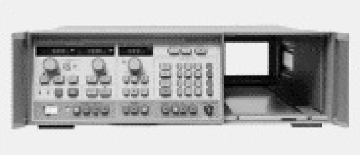
Finally, you may see the Marconi 6600. This is quite large and very heavy and like the HP 8690 has a range of BWO plug-ins up through the mm-wave bands. Getting a bit old now but if the unit is working then still worth buying.



Interconnecting these with your other test equipment can be a bit of a nightmare, especially as you are likely to want to change things around frequently, unlike a professional lab. No, I don't have a clever answer to this but since the connections are mainly BNC it isn't too much of a chore (as I've just rebuilt the shelving for the test gear I can tell you that not having it all against a wall is a big help) and in some cases you can use BNC T pieces to feed two items a once.. Those with time on their hands, having completed the Telegraph crossword, may like to ponder over a nice relay matrix to solve the problem.

If you are starting off then you probably want to use the sweeper with a scope and a detector.

The problem is the output from the sweeper is almost always 0 to +10V and a few scopes want +/- ve on the X axis. Not a big problem as you can convert it with an op-amp on dual rail supplies. You can also use an op-amp if you need a greater voltage swing. You can use an op-amp to work the other way around as the popular 141 spectrum analyser sweep output has a +/- 5V swing – yes you can use the first LO as a limited range sweeper.



Next up are the scalar network analysers. The two I know of currently in circulation are the Wiltron 560 and the Marconi 6500. These differ in that the Wiltron requires a sweep signal input and the Marconi generates the sweep signal. Connecting the Wiltron is simply a matter of following the information on the back panel but the 6500 won't directly connect to an HP 8620 BNC's, although it will the 8350. Russ, G4PBP, has successfully connected the two by using the sweep input connection on the 50 way accessory programming connector. Pin 28 is the sweep input and Pin 43 is ground. Russ brought these out to a SMC connector on top of the back panel, taking care to ensure that the socket clears the top cover. Connect the SMC to the 0-10V BNC on the 6500 and follow the programming cards in the 6500.

Finally, there is the 8410 vector network analyser. Whilst this is a very specialist instrument it is extremely versatile – tuning multistage filters up is even more of a doddle than using a scalar network analyser and, of course, it is *the* tool for antenna measurements. I'll take some time in another issue to describe the differences between the two types.

The 8410B/C works well with both the 8620C and 8350 sweeper but really needs a programming connector cable to carry some additional control signals – the 8410A doesn't have this capability. Of course, these never come with the units. Fortunately there are only three connections so it is easy to make one up.

Connection	**8410B/C J17**	**8620 J2**	**8350 J13**
Stop Sweep Pulse	Pin 7	Pin 34	18 and 20
Seq. Sweep Trigger	Pin 1	Pin 26	24
External Trigger	Pin 9	Pin 50	n/c
Ground	Pin 11	Pin 43	19

Simple Microwave System Benchmarking

Uffe Lindhardt PA5DD

This article describes a simple way to test the receive performance of your microwave system. It requires that you are able to receive a signal from a known source from within 0-25 km (e.g. a beacon or a neighbouring amateur).

The method relies on free space propagation theory, so it is important to have an unobstructed path to the source. The method does not include any compensations for refraction, reflection, atmospheric losses or ground gain, hence results should be used with caution. Neither does it take account of antenna illumination efficiency and receiver performance. The idea is to keep the calculations as simple and clean as possible. Afterwards the figures can be seen in the light of the omissions.

To further increase transparency and simplicity the units for antenna gain has been adapted to the free space propagation formula, so that the frequency independence of the spreading of energy over the path becomes apparent. Therefore the effect of the transmitter antenna is characterised in dBi and the receive antenna in aperture. This approach and the use of dB as unit makes the calculations a matter of adding together some figures. Note that anything but the TX EIRP are frequency independent, which is convenient when comparing different bands, where the same multi band parabolic dish is used. The antenna aperture is approximated to be equal to the area of the parabolic dish used.

To perform the benchmarking you need to be able to determine the signal-to-noise (S/N) ratio of the incoming signal. This is most easily done using a adjustable attenuator in the IF line (e.g. 144 or 432 MHz) of the microwave transverter. You should preferably have so much IF gain that the noise without attenuation produces a S-meter reading. After having determined that S-meter reading (with fast AGC), you tune to the signal for a maximum reading, afterwards you add attenuation until a reading equal to the noise reading is achieved. The S/N should be at least 20dB for this method to work.

Now you are ready to fill in a table like the one overleaf . You need to know the EIRP of the source signal, the distance to the source, the radius of the parabolic dish used and the bandwidth used for the measurements (normally 3kHz). Start at the top, and calculate your way to the bottom. When comparing the calculated and measured figures, be aware that at least the first 7 - 10dB of difference is accounted for by the omissions in the calculations. The antenna illumination efficiency accounts for 5 - 6 dB, receiver noise figure and feeder loss for another 2 -3 dB. The rest can be attributed to propagation factors, faults in the system or inaccuracy of the data used (e.g. the true EIRP of the source).

Conversion between dBi and antenna aperture can be made using the following formulas:

Antenna gain [dBi] = [Aperture in dB/m^2] + 10 * Log ((4 * ?) / [wavelength in meters] ^{2})

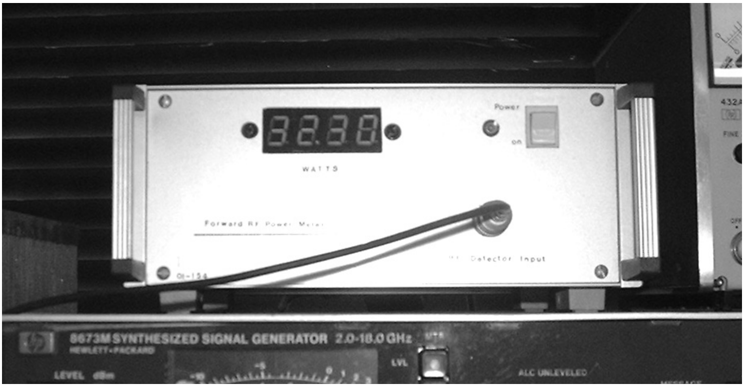
Calculations		**Unit**
TX EIRP	10 * Log [power in mW] + TX antenna gain (dBi)	dBm
Path gain	10 * Log (4 * ? * [distance in meters]^2)	dB
=		
Power density at RX antenna		dBm/m^2
RX antenna aperture (ideal)	10 * Log (? * [parabola radius in meters]^2)	dB/m^2
=		
Power at RX input		dBm
Noise floor @290K, 0dB NF	10 * Log [bandwidth in Hz] - 174 dBm/Hz	dBm
<>		
Ideal free space S/N		dB
Measured S/N		dB
<>		
Difference ideal to measured		dB

Shown on the next page, is my own benchmarking using PI7EHG at Schiphol airport. The distance is 25km, and is very close to line-of-sight because the beacons are placed 90 meters higher than the receive antennas. I use the benchmarking to get a picture of absolute performance, but more importantly to track any changes in system performance over time.

PA5DD JO22IC22	**3400 MHz**	**5760 MHz**	**10368 MHz**	**24192 MHz**
PI7EHG (JO22JH14) EIRP	+ 39dBm	+ 41.5dBm	+ 32dBm	+ 26dBm
Path gain (25km)	- 99dB	-99 dB	- 99dB	- 99dB
=				
Power density at RX antenna	- 60dBm/m²	-57.5dBm/m²	- 67dBm/m²	- 73dBm/m²
RX antenna aperture (ideal)	- 4dB/m²	- 4dB/m²	- 4dB/m²	- 7.5dB/m²
=				
Power at RX input	- 64dBm	-61.5dBm	- 71dBm	- 80.5dBm
Noise floor @3kHz,290K,0dB NF	- 139dBm	- 139dBm	- 139dBm	- 139dBm
<>				
Ideal free space S/N @3kHz	75dB	77.5dB	68dB	58.5dB
Measured S/N @3kHz	57dB	52dB	56dB	36dB
<>				
Difference ideal to measured	18dB	25.5dB	12dB	22.5dB

Digital Display for a Microwave Diode Detector

John Share, G3OKA



A common technique used to determine voltages at Radio Frequencies is to employ a diode detector circuit designed specifically for the frequencies and power levels of particular interest. The resulting dc voltage is proportional to the RF voltage, however this relationship is not linear due to diodes having a "Square Law Characteristic". Custom scales can be made for Analogue Panel meters but some form of linear conversion is necessary for a digital display.

Analogue Conversion

The square root of a value is the Antilog of half its Logarithm. Obtaining a square root using analogue circuitry requires converting the input to its logarithm, dividing by two, and then converting to the antilog to obtain the result.

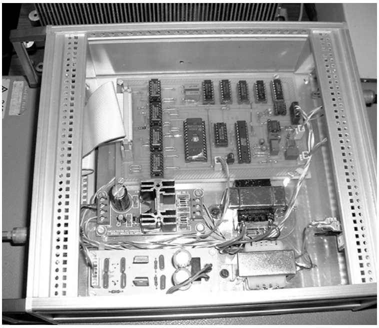
E(out)= Antilog(0.5 * Log(E(in))

Integrated circuit manufacturers have devices that perform Log and Antilog functions, they are expensive and are not too readily available. The building blocks to create these functions are well known (e.g. Clayton, Operational Amplifiers) and it was initially envisaged that a circuit could be constructed and the linear output displayed on a Digital Panel Meter module.

The design and construction of Log or Antilog Amplifiers should not be undertaken without considerable caution. Typically transistor base / emitter junctions are employed to mimic the Log function, these junctions are very unstable due to thermal drift. A prototype built to the Clayton design proved to be virtually impossible to calibrate due to thermal instability. It was quickly rejected as unusable and merely served to underline the justification of the cost of commercial devices.

Digital Conversion

An alternative is to use digital techniques by converting the input signal to a binary number, processing this value in software and displaying the result. Such a task can be readily undertaken by a PC however this is



excessive and a minimal system can be designed. All the input values can be predicted and a look up table can provide a direct output without the need for software processing.

It is necessary to decide on resolution in order to determine how many bits of data are required in the system. Conveniently 7 segment display drivers require four data lines of input, EPROM chips have 8 lines of data output hence they can function two decades of display per byte, two bytes will function four decades of display.

The specific requirement was to display power levels to a maximum of 40 Watts with a resolution of better than 0.1 W, i.e. four decades of display. Digitising the input signal (2.7 volts = 40 Watts) to a comparable resolution required particular consideration of the input at small values where the differential is least pronounced. For correct resolution across the full range it is necessary to use a minimum of ten bits of Analogue to Digital Conversion.

The input signal is digitised using an ADS7806 Analogue to Digital Converter, the output values are connected to the address inputs of the 27c64 EPROM that has been loaded with the 4 digit output value for every input value. The data outputs drive four 7 segment display latch / encoders (CD4511) that in turn power the digital display.

The ADS7806 is configured for parallel data output and places a digitised value on its output pins in two bytes as selected by pin 21. The lower byte consists of only the four least significant bits and these are latched into the 73LS173, the upper byte remains present whilst the upper byte is selected. In this manner the 12 bits of the EPROM address are obtained from the two bytes of data.

It is possible to configure the ADC for a multiplicity of input ranges but for compatibility with other equipment it was desirable to have a 0 to +10 volt input range. (The 0-10v Output shown on the block diagram is taken to a PC where it is digitised for system control.) Selecting the Complementary Offset Binary (COB) data format had the advantage of the most significant bit being asserted when the input went negative. The EPROM was loaded so that all addresses with A11 high were decoded as "0000" to the display. Data in the EPROM was generated by a program written in QB45, this calculated all the data points and saved them on a floppy disc as a text file (Data.txt) that could be viewed in Microsoft Word. A Softy EPROM Programmer was used during development and a compatible file created from this QB45 program was transferred by serial connection at 9k6 baud.

The EPROM address line A12 selects the upper and lower pairs of displays thus with A12 low the display data for the two upper digits are selected and with A12 high the lower two digits are selected.

Timing strobes are generated using a binary counter (73LS93) and an 8 line decoder (74LS138). On the leading edge of t0 a 4 us convert pulse is sent to the ADC which will complete within 16 us. The timing oscillator is considerably slower and t1 occurs much later generating the strobe pulse to the 74LS173. Binary counter output "B" and “C” are “NORed” to select the upper byte and output “C” switches the EPROM between the upper and lower digits. The display latches are loaded at t3 and t6.

Calibration and Setup

Zero offset and gain adjustment trimmers are provided for the input amplifier. The OP07CN has excellent thermal characteristics and does not compromise the system. The specific unit was required to display 40.00 Watts for an input signal of 2.700 Volts. This was derived from a bench power supply and the only adjustment was to set the gain of the input amplifier for an output of 10.00 volts.

There are a number of external adjustments that can be made to the ADS7806 however these proved unnecessary apart from trimming the zero offset. It is essential to take into account the input amplifier when checking the display for given inputs. Display = 40 (SQR (ADC(in)/10)).
As a guide 1.35 volts applied to the input connector equals 5.00 volts input to the ADC and displays 28.28 Watts.

In Operation

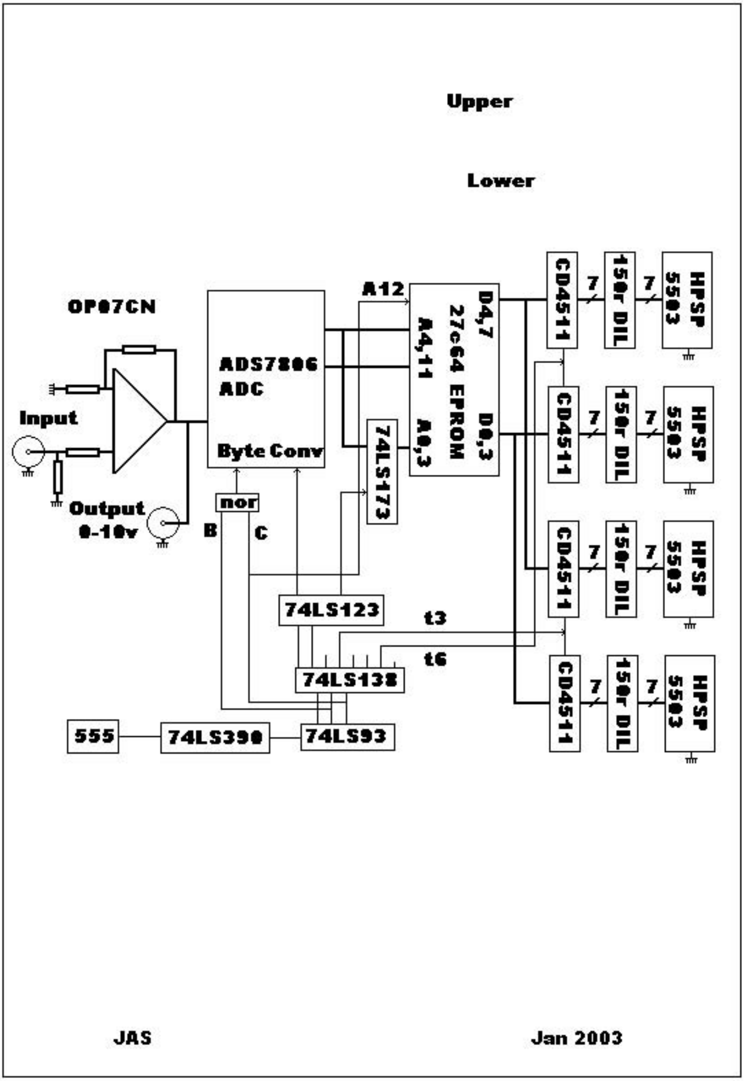
The most remarkable feature of this unit is its display stability, when driven by a test signal the output remained locked for hours on end. Digitising invariably results in LSB dither however the choice of ADC resolution appears to have circumvented this problem and the display LSB is stable, resolution exceeds the original requirement.

Acknowledgements

This unit was designed and built as part of an 18 GHz Microwave system to investigate magnetisation in rock samples at the Geomagnetism Laboratory, University of Liverpool.

EPROM PROGRAM FOR THE UNIT

```
Start:
OPEN "A:DATA.TXT" FOR OUTPUT AS #1
FOR adcout = 0 TO 2047
Vin = adcout / 2048
a = INT(SQR(Vin) * 1024)
Watts = 40 * a / 1024
hexadc$ = HEX$(adcout)
hexstr$ = HEX$(a)
hexadcupper$ = HEX$(adcout + 8192)
MSB = INT(Watts)
LSB = INT(100 * (Watts - MSB))
PRINT #1, hexadc$; " "; MSB; " ";
hexadcupper$; " "; LSB; " ",
NEXT adcout
CLOSE #
```


Upper
Lower
OP07CN
Input
Output
0-10v
ADS7806
ADC
Byte Conv
A12
D4,7
27c64 EPROM
A4,11
A0,3
D0,3
74LS173
nor
B
C
74LS123
t3
t6
74LS138
555
74LS390
74LS93
CD4511
7
150r DIL
7
HPSP
5503
CD4511
7
150r DIL
7
HPSP
5503
CD4511
7
150r DIL
7
HPSP
5503
CD4511
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Jan 2003

Directional Couplers used for VSWR and Power Measurement

Richard T. Knadle, K2RIW

INTRODUCTION

Over the years, I have heard many engineers and some smart amateurs, express opinions that reflect a considerable misunderstanding about the operation of Directional Couplers, and how to properly use them in the measurement of Voltage Standing Wave Ratio (VSWR), and power. This memo is intended to give some basic information that may help. At first, the average electronic technologist is mystified by at least two of the concepts of how RF behaves within transmission line structures:

(1) The concept of a Directional Coupler (DC); the idea that it favours a signal that flows in one direction, yet rejects (at least partially) a signal that flows in another direction seems (to them) to be in violation of some basic laws -- like the Law of Reciprocity.

(2) On top of this, many technologists have great difficulty believing that a normal transmission line, completely keeps separate, the signals that flow in the two directions on that line, even if those two signals originally came from the same source.

I believe that both of these principles must be absorbed (and understood), if meaningful DC measurements are to be properly executed and believed. Here are my recommended procedures, with some partial explanations of what is taking place at each step.

A DIRECTIONAL COUPLER USED IN A VSWR OR POWER MEASUREMENT PROCEDURE

(I) DIRECTIONAL COUPLER CALIBRATION -- The first step in this procedure is to establish the quality of the Directional Coupler (DC) that you are about to use. I don't care if the label on the DC says it is a "Cadillac" or "Rolls Royce" brand, and the calibration sticker says it is traceable to "The Bureau of Standards" with and accuracy of 0.01dB; you still have to confirm that it is good working order right NOW. It is possible that the DC was thrown onto a concrete floor yesterday, and the internal termination may have been shattered. If that had happened, it could loose almost all of it's directional characteristics -- it's "Directivity."

The confirmation requirement is similar to the proper use of an Ohm Meter. Notice that a good technologist will always short the two leads together; and the Ohm meter had better read a small fraction of an ohm, before the technologist will proceed with the next measurement.

Similarly, a prudent technologist will measure the Directivity of the DC he is about to use. It is also useful to know that sometimes the DC can be used far outside the frequency range it was designed for, as long as the principle of operation is somewhat understood, and a calibration at the present frequency is performed. Here is the Directivity Confirmation procedure.

DIRECTIVITY CONFIRMATION -- Unfortunately, the Directivity Confirmation procedure requires a known good termination (dummy load), and the procedure will have an accuracy that rarely is much better than the quality of that termination being used. First apply an RF signal to the DC "input" port, with a known good termination connected to the "output" port. Position the DC so that it favours the Forward flowing signal. Place a power-measuring device at the directional port. This can be a Power Meter, Spectrum Analyzer, calibrated Crystal Detector, Scalar Network Analyzer, etc. Measure (and record) the DC's response to the forward-flowing signal (in dBm units). If, for instance, you are using a known Directional Coupler (DC) with a -10dB Coupling Coefficient, the measured power should be nearly 10 dB weaker that the power that's being applied from the signal generator. By the way, "dBm" means Decibels of signal strength with reference to a 1 milliwatt signal.

Next, reverse the DC "input" and "output" ports, and repeat (and record) the previous measurement. The difference in the two readings indicates the Directivity. For instance; if a 0.0dBm signal generator is applied to a 10dB coupler, and it measured -10dBm during the Forward Measurement, and -30dBm during the Reverse measurement, that would indicate a Directivity of 20dB (the difference in the readings). A DC of "Good" quality will show a directivity of 2 dB; that is,
the apparent reflection from the termination will appear to be -20dB (an apparent VSWR of 1.22:1), even if the termination is a perfect 50 ohm resistance at the present frequency. An "Excellent" DC will show a Directivity of 30dB (an apparent VSWR of 1.065:1), and there are Instrumentation-type DCs that can display a Directivity of over 50dB (an apparent VSWR of 1.006:1).

More on this later; there are ways of improving your DC's Directivity. Simplistically, you could say that a DC that displays a Directivity of 20dB will not be able to easily resolve the Reflection Coefficient from an unknown load of better than about -20dB and there are ways to get around this. Depending on how well your DC is internally balanced, the finite Directivity (-20dB for instance) represents the degree of response it has to a signal that is flowing in the wrong direction -- this is really it's degree of imbalance. A modern Network Analyzer uses a complicated "12 point" calibration procedure to drastically improve the accuracy of a Reflection measurement it makes with it's "only Good quality" Directional Couplers.

ALTERNATE CALIBRATION PROCEDURE -- There is an alternate Calibration Procedure that does not require the inconvenience of reversing the DC to measure it's Directivity. This is to recognize that a good Short (or Open) circuit has a Reflection Coefficient of nearly -0.0dB. In this method, first measure (and record) the apparent reflected power from a Short (or Open) termination, then place the Known Good Termination on the "output" port of the DC and repeat the measurement. The difference (in dB) between the two measurements represents the DC's Directivity. When using SMA or type N connectors at 10GHz (and below), an "Open Circuit" will have Reflection Coefficient of nearly -0.0dB, and is a good calibration "short/open termination." However, if you're using a Wave Guide (WG) type DC, an open circuited WG flange makes a pretty good transmitting antenna, with a VSWR of about 1.5:1 (reflection coefficient of about -12.9dB). Therefore, don't use this as a high reflection termination. Instead, place a sheet of metal (tightly) across the WG flange as the high reflection termination.

SIGNAL GENERATOR VSWR -- There is an additional danger to the alternate calibration procedure. It is vulnerable to the VSWR of the signal generator. I would only use this procedure if there was a 10dB (or greater) pad between the signal generator and the DC. Without that pad, the reflected signal could re-reflect from the signal generator and cause a confusing reading. The signal generator reflected voltage can add to the incident voltage and create an apparent signal source that would appear as much as 6dB greater (or more) in magnitude -- but only during the short/open portion of the test. Also, if the DUT happens to have a rather high VSWR (reflection of greater than say 20dB), I again would recommend the use of a 10 dB pad at the signal generator.

(II) THE UNKNOWN MEASUREMENT -- Once you have confirmed that your DC is performing properly, it is time to place the Unknown Circuit (the Device Under Test [DUT]) on your DC to measure, and tune, it's Reflection Coefficient. The DUT-reflected signal can then be translated into VSWR by using a look-up table or by performing a two step calculation.
Step (1): Convert the reflection coefficient (in dB) into a reflection Voltage, which is usually represented by the Greek letter Rho.
Step (2): Convert the Rho magnitude into VSWR.
(1) Rho = ALOG(-dB/20)
(2) VSWR = (1 + Rho) / (1 - Rho)
Where:
ALOG = Anti-LOG, or 10^(-dB/20)
Rho = |Absolute Value| of the Reflection Coefficient (as a Voltage).
The final dB of Reflection Coefficient in the numerator must be a negative number that's then divided by 20 and raised to the power of ten in formula (1). At first, some technologists will understand that the dB value is negative dBs, they place it into the formula that has another negative sign in it, that converts it to a positive value (+) and they come up with answers that are crazy.

CHEAP AND BROAD -- The beauty of using a Directional Coupler (DC) in VSWR measurement is that, generally, they are rather inexpensive, and they are rather broadband, therefore a swept frequency measurement is possible if your power detector is a fast acting one, such as a calibrated Crystal Detector (and oscilloscope), a Spectrum Analyzer, or a Scalar Network Analyzer (SNA). As you tune your DUT, it is nice to know that you are tuning for a broadband match, as opposed to an impedance match that is only effective across a narrow frequency range.

(**III) DC ALTERNATES --** There are a large number of devices that can serve as the Directional Coupler (DC). They have such names as Quadrature Hybrid, 90 Degree Hybrid, Branch Hybrid, Branch Coupler, Magic T, Ring Hybrid, Zero-180 Degree Hybrid, Wave Guide Broad Wall Coupler, Wave Guide Narrow Wall Coupler, Wave Guide Beth Hole Coupler, etc. The one kind of hybrid that
can't be used this way is a Wilkinson Half Hybrid, or Zero Degree Hybrid.

(IV) DC EXTENDED FREQUENCY RANGE -- Few technologists know that a well constructed Directional Coupler (DC) has an operational frequency range that extends many octaves in the lower frequency direction. For instance, if you plotted the Forward Response of a DC that's rated for operation from 1 to 2GHz, you would find that it has useful operation all the way down to 10MHz (and probably below). The only thing that changes is its frequency flatness, and the Coupling Coefficient decreases -- but that can be a considerable advantage. Here is what's happening:

(A) A TEM-type (non Wave Guide type) Directional Coupler has its greatest coupling at the frequency where the internal coupling section is 1/4 wave long. Above (and below) that frequency the response falls off in a very predictable manner -- it's a SINE wave of amplitude. In other words, if I was sweeping that DC that's rated for 1 to 2GHz, and I plotted the Foreword absolute Voltage response versus frequency at the Coupled Port, the resultant plot would look like a rectified SINE wave, with the horizontal axis being frequency (instead of time). There would be a zero response a zero MHz, a broad peak near 1.5GHz, a second zero near 3GHz, a second broad peak near 4.5GHz, etc. Unfortunately, a DC only has Directivity at the 1/4 wavelength frequency region and at lower frequencies -- but that still leaves many octaves of useful operation.

(B) That predictable response outside of the rated frequency range has turned into an advantage for me on many occasions, here are some examples:

(1) For my first published article, "A Stripline Amplifier/Tripler for 144 and 432MHz", Ham Radio, February, 1970,

I needed to test the power output, and harmonic content, of the 144MHz section and the 432MHz tripler section of that 4CX250B amplifier. I needed a 300 watt frequency-indicating power meter, that I didn't have. A Spectrum Analyzer (SA) can do the job, but it can't tolerate the 300 watts. If I had a -30dB DC, the coupled power would be 0.3 watts and the SA could easily make the measurements. But, my company's Instrumentation Department said they didn't have a -30dB DC at that frequency range, and non of their DCs could tolerate 300 watts. I studied what they had and found a solution. They had a Narda -10dB type N Directional Coupler rated for 8 to 12GHz and 1 watt maximum. I reasoned that the coupling section was 1/4 wave long (90 degrees in phase length) at 10GHz, the centre of it's frequency range. I then divided 144MHz by 10GHz, multiplied by 90 degrees, and reasoned that the coupling section was only 1.296 degrees long at 144MHz. The SIN of 1.296 degrees is 0.02262. Since this is a voltage response I took 20*LOG(0.02262) = -32.9dB.

That means that the coupled response at 144 MHz would be -32.9 dB (weaker) than at 10GHz, where it was a -1 dB coupler. Therefore it is a -42.9 dB coupler at 144MHz. I calibrated it at 144MHz and found it to be a -43.1 dB coupler -- close enough. And, since the internal coupled line is isolated from the main line by -43.1dB, that means that the internal 50 ohm termination would never see more than 0.015 watts when I applied 300 watts of 144MHz signal to the coupler. I similarly calibrated it at the harmonic frequencies, applied the 300 watts to it, it worked like a charm, I made all the measurements this way and they appeared in the article.

In the low frequency area of a coupler's response (near 0 degrees of a SIN function) the response is almost a straight-line response that falls off at -6dB per octave (-20dB per decade) as you go down in frequency. Therefore the "-43.1dB coupler" I used at 144MHz would be a -63.1dB coupler at 14.4MHz. As you are about to see, Directional Power Meters use this principle.

(V) BIRD-TYPE POWER METERS -- It is interesting to note that the slug of a Bird Power Meter is also a less than 1/4 wave section of a Directional Coupler. The Bird slug achieves frequency flatness across it's rated frequency range by using a rectifier circuit that has a low-pass filter action that rises at 6dB per octave as you go down in frequency.

Each slug also has a finite Directivity, depending on how well it was balanced and calibrated at your favourite frequency. Therefore, be careful about falling into the trap of using a high power slug to measure the forward power of your 1 kW XMTR, and then switching to a low power slug to measure a very low VSWR. Your antenna may be perfect, and have no reflected power (voltage), but the slug's approximate 20dB of Directivity could show an apparent antenna reflection of -20dB (10 watts). That could lead you into believing that the antenna VSWR was 1.22:1.

(VI) COUPLER IMPROVEMENT TECHNIQUES --
As the above material shows, a DC that has less than ideal Directivity is really displaying a slight imbalance that causes it to slightly respond to the signal that is flowing in the wrong direction on the main line of the coupler. There are many ways of improving the DC's balance.

(1) Internally, you could re-adjust the accuracy of it's termination, or you could add a small gimmick capacitor in the correct location to improve the Directivity balance.

(2) An even better way is to use a Double Slug Tuner, a Double Stub Tuner, or a Wave Guide E - H Tuner. If you have a known good termination, you can assume that it has perfect absorption and essentially no reflection. You then place the tuner between the DC and the good termination, and adjust it until the DC shows no reflected power from the termination. You then leave the tuner connected to the same port of the DC, while you proceed with the VSWR or power measurements.

When you were adjusting the tuner for a null in the DC's Reflection response, you were really creating a second small reflected signal that was equal in amplitude and 180 degrees out of phase at the DC coupled port. That created the improved balance and made the DC nearly ideal, at that frequency. The bandwidth of this DC correction technique is dependent on the amount of correction that was required. When in doubt, recheck the balance at the next frequency.

(VII) TRANSMISSION LINE DIRECTIONALITY --
When I tell a technologist that a transmission line will keep the two signals completely separate, that flow in opposite directions on a transmission line, they often don't believe it -- particularly if the two signals came from the same source. There are many RF tests that could be performed to prove this, but I have discovered that a well informed sceptical person can always come up with an alternate explanation that supports their point of view.

I have found that the best way is to use visual experiments.

(1) A pool of water is really a radial transmission medium. If I drop a pebble at the North end of the pool, waves will travel to the South. Similarly, a pebble dropped into the South end will create waves that travel to the North. If I drop pebbles at both ends of the pool, the waves will meet at the middle, and pass right through each other with no interference, as long as the waves are kept small enough (use the linear region of wave amplitude -- no white caps).

(2) I can tap the 1/4 inch guy wire on my 200 foot Rohn-55 tower and watch the wave travel up the guy wire, strike the tower, reverse in polarity, and propagate back down to me (it hit a "short circuit"). I can wait until the wave has struck the tower, and started

back to me, then I can strike the wire again (with any polarity) to start a second wave going up the guy wire. As the two waves meet in the centre, they pass right through each other with no interference, as long as the waves are small enough that I don't get into non-linear stretch (deflection) of the steel. (3) I say that most linear transmission mediums obey this property -- even RF in free space. Those waves that meet in free space pass through each other with no real interference. When you move your Handy Talky Radio around a room that is reflective, you will find what you think are signal nulls. This is because you are using an antenna that has no Directivity and it is responding to (adding together, voltage-wise) at least two waves that are out of phase. Similarly, the probe that is used on a Slotted Line VSWR setup has no directivity, and it displays the Standing Wave Ratio that is caused by the signals it picks up that flow in both directions through the Slotted Line. Those two signals were completely independent up until the time they were combined on that non-directive probe wire and then, for the first time, they interfered with each other to generate the effect we call "a standing wave." This measurement technique has become the classic way of specifying the Reflection Coefficient of an RF device -- its VSWR.

(VIII) LETS DO AWAY WITH VSWR -- If you took the directional probe from the slug of a Bird Power Meter and operated it on that Slotted Line, you would discover that the Standing Wave has disappeared, and you could now independently measure the amount of power (or voltage) that is flowing in each direction (by reversing the slug) -- that's really what you wanted to know in the first place.

In the past, that Slotted Line measurement was the only way you could conveniently measure the reflected voltage -- by using an interferometry technique to indirectly measure it as VSWR. It really is time that we abandon "VSWR measurements" because we don't do it that way any more. We should only discuss the Reflection Coefficient -- in watts ratio, volts ratio or dB ratio (choose your favourite units), because we now directly measure the reflected signal. We RF mavens seem to spend half our lives converting back and forth between VSWR, Voltage Reflection Coefficient (S11, S22) or Power Reflection Coefficient, just so that we can communicate with a technologist (or the data sheet) that uses the other system of units.

"VSWR" is now a "coded message". It's really time that we "Break the Code" or stop using that code when we're buying components or training the new RF recruits. I'll admit that we will have to keep mentioning it to students, for historic reasons.

(IX) TROMBONE IMPROVEMENT
I'll warn you that these last three paragraphs will only be appreciated by a person with a rather exacting-type of personality!

Once you accept the fact that RF power can independently flow in two directions on a transmission line, you then realize that changing the length of a lossless transmission line (of the same impedance) does not change the Reflection Coefficient; thus it doesn't change the true VSWR of your antenna. However, if the Directional Coupler (DC) device you're using (coupler or a Bird) has less than ideal Directivity, than the Reflection Coefficient, and VSWR, will appear to change. This is because there is a small amount of Forward-flowing signal (I'll call it the Leakage Signal) that's mistakenly Being picked up by your coupling device, that beats against the real Reflected Signal that your coupler is now measuring (from your antenna, for instance). As you change the length of the transmission line (with a Trombone Line), the two signals go in and out of phase with each other. This will show up as a cyclicity of the apparent Reflected Signal Power, as the Trombone is operated. This assumes that your trombone can move more than one half wavelength at your frequency – you're not going to do this at 80 meters! Although on 80 meters you could insert fixed lengths of low loss cable (of the same impedance) to get the same effect.

Knowing the operation of the system, and its shortcomings, can allow you to gain higher accuracy in the Reflection Coefficient measurement. A perfect DC or Bird would show no change in reading as the Trombone (on the antenna side) is operated. The magnitude of the "ripple" is an interferometry effect that is telling you exactly how strong is the Leakage Signal into your coupling device. Once you know the strength of the Leakage, you can subtract it out of your measurement. This is exactly the accuracy improvement procedure that is done in the microprocessor of a modern Network Analyzer.

You can convert the Ripple into a Leakage Magnitude by using the following formulae:

Leakage Voltage = (a - 1) / (a + 1).
Leakage Voltage (dB) = 20*LOG[(a - 1) / (a + 1)].
a = ALOG[Ripple / 20].

Where:

Ripple is expressed in Peak-to-Peak dBs, a positive number.
LOG is calculated in base 10.
ALOG is the Anti-Log, or 10^(Ripple / 20).

"a" must be a positive number, greater than 1.

Here is a measurement example. Assume I'm measuring the Reflection Coefficient of my UHF antenna system and my DC says that the Reflection is around -19.5dB. As I operate the Trombone after the Coupler, I see a Peak reading of -19dB, and a valley reading of -21dB. That's a Peak-to-Peak reading of 2dB. The formula tells me that my Leakage Signal is 0.1146, or -18.81dB (weaker) than the Peak and Valley measurements I have made.

That relative Leakage voltage was in-phase at the -19dB reading, and out-of-phase at the -21dB reading. I can choose to subtract the voltage from the -19dB, or add it to the -21dB reading. This relative voltage will thus be 1.1146, or 0.9954 (as a voltage), and I can take 20*LOG of these voltages. Thus, I can either add 0.94dB (in absolute terms) to the -19 reading, or subtract 1.06dB (in absolute terms) from the -21 dB reading. In either case the corrected reading will be an antenna Reflection Coefficient of -19.94dB.

I hope this information is useful to those who could read this far. Feel free to correct the mistakes!

Chapter 13

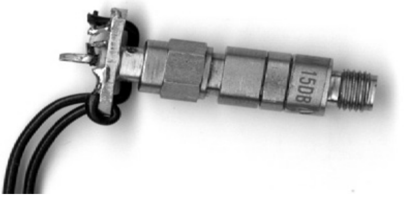
Noise

A Noise Source and its Calibration

David Wrigley, G6GXK

The news that an Agilent Noise Measuring Equipment was to be provided at RAL 2001 in April this year was an opportunity too good to miss.

Noise measurement was a thing I had not been involved in and, having by chance just completed a homebrew SLNA preamp for my 10GHz transverter, now seemed as good a time to start as any. A quick check of the info around soon revealed the excellent articles by Paul Wade (W1GHZ) previously published in the Microwave Newsletter for November and December 1997 but available online www.qsi.net/n1bwt/noise99.pdf - (Ref. 1). It was decided to build three noise heads using the most likely devices available in the UK and see how well they performed on the day.



Basics

The noise head works by using the random noise produced by the avalanche breakdown region of a semiconductor junction. In order to achieve this, a reverse voltage has to be applied to the junction. Clearly the current under these conditions needs to be limited or the junction will be destroyed, and this limiting is achieved by means a series resistor. However the precise current which gives the best noise output at any frequency or the flattest output across all frequencies has to be determined by experiment and there is usually a potentiometer provided for that purpose. In choosing a suitable device, the base-emitter junction of a microwave npn device seems to work well although there are advantages in reliability and stability in using a specially designed diode such as those produced by Noisecom. These latter devices have been designed to minimise flutter effects and instability over time.

History

With only two weeks to go to the RAL event and other priorities competing, it was decided to buy a selection of likely devices from Farnell and check these out prior to the proper tests at RAL. It was also decided to build a skeleton assembly type source from Paul Wade's article and also two samples of the later Teflon PCB type using some available Rogers material.

All of them worked but, as luck would have it, the skeleton type proved to have to flattest frequency response from 28MHz to 10GHz. The results are shown below. I must add at this point that these and many other measurements were taken with Ian (G3SEK) operating the controls for much of the day at RAL and it is to Ian and Dave Stockton of Agilent we owe our thanks for the operation and provision of such a great piece of test gear.

The noise head had an external 15dB SMA attenuator attached to it throughout the tests and I figured that it would best to keep that attenuator with it.

Construction:

You will almost certainly require some sort of magnifier to assemble this head. I use a headband binocular type, which is "hands free" and doesn't require the watchmakers ability to screw up the eye to hold an eyeglass. You will need a fairly powerful soldering iron – I used a 40 Watt temperature controlled Weller with a pointed tip.

The construction is based on a two hole, female SMA chassis mounting connector. The procedure is as follows:

1. Screw a male connector onto the female connector so that it can be gripped in a vice rear side up.

2. As close as possible to the Teflon form a pool of solder and then drop a 1pF SMD capacitor in it –end on. It will stay that way because of the way the solder surface pulls it. I then tilted the cap towards the central conductor of the connector, because I wanted to make sure that the transistor would fit. Take care only to try to move the capacitor whilst the solder is molten otherwise it will fracture near the termination. – (Been there – done that)

3. About a chip length back from the 1pF, mount a 1nF capacitor vertically in another pool of solder. This will be near the fixing hole of the connector – so make sure that it is to one side of it to keep the hole clear.

4. With tweezers carefully mount a 47R (marked 470) SMD resistor across the top of the two capacitors.

5. With tweezers carefully mount the base emitter junction of the BFS520 transistor between the top of the 1pF and the central conductor. If you have done that right , the markings of the transistor will be uppermost and with the collector away from you, the base (LHS) will be connected to the central conductor.

6. Solder a wire near the other side of the Teflon as a ground connection, but don't thread it through the hole first – unless you are using Teflon insulated wire.

7. Thread the ground wire through the hole and thread another wire for the supply current through both holes as shown in the photos. This wire can then be soldered to the top of the 1nF where it connects to the 47R. Threading the wires in this way will minimise the possibility of the wire breaking off or worse, pulling off the termination of a component.

And that's it, apart from putting a wire ended 1K fixed resistor (limits the max current) and a 10K pot in series with the live lead, and also not forgetting to add a BNC connector on the end of the wire, for the supply voltage.

Supply voltage differences

The supply voltage to the head when driven from the Agilent Equipment is 28 Volts. When I built my PANFI (Precision Automatic Noise Figure Indicator) I decided to use 15 Volts, because this was the voltage used by the higher frequency heads manufactured by Noisecom – and one never knows I might be lucky enough to come across one.

So I had a noise head calibrated at 28 Volts, with an external series resistor of 0.98K + 7.9K (total resistance 0.98K + 7.9K + 47R + 50 R = 8.977, near enough to 9K total) which now needed to have the series resistor changed to enable it to work on 15 Volts. In order to work this out I needed to check the Avalanche breakdown voltage at the current at which it was calibrated. So I supplied 28 Volts and measured it – this turned out to be 5.28 Volts (measured across the diode), and a calculated current of 2.53mA. The total series resistance required for 15 Volts at 2.53mA is 3.84K. The 47R series resistor and the 50R attenuator input resistance need to be subtracted from this to give the required external resistor 3.74K. A combination of a 3.9K and a 180K in

Frequency MHz	Noise Level dB
28	-7.2
50	-6.8
70	-6.7
144	-6.6
432	-6.6
1296	-6.5
2320	-4.0
3450	-9.0
10368	-6.1

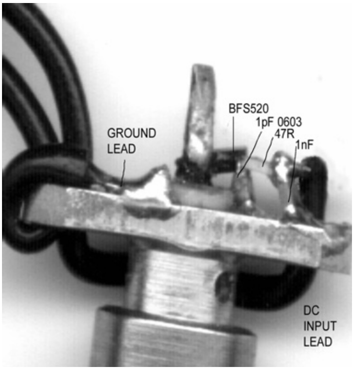
parallel was used to replace the external resistor and potentiometer. This had the advantage of no potentiometer and therefore no accidental way of it changing.

Conclusions

Well, some checks have been made using the head and a home made PANFI (based on ref.1) and now, for the first time, I can see the significant difference in noise figure between a straight FT290 and one with a Mutek front end.

There was also found to be a slight benefit in using the Mutek front end in preference to the pre-amp in the 100W PA.

I then decided to try the Noise head at 10GHz . Here unfortunately, it demonstrated that my attempts at designing my own SLNA for 10GHz was less successful than the original G3WDG transverter, which has a NF of between 2 and 3dB. I did try tabbing the lines to improve the matching but try as I could, I could not bring it to the 1dB NF which I believed was possible. I therefore removed the preamp in order to carry out further tests and most probably I will rebuild it later this year.



This noise figure measuring system has certainly proved to be a most useful tool and has removed some of guesswork when tuning up receivers. At least I can be confident that I have got my rigs to the best they can do at the time and have revealed the weaknesses which can be the focus of future improvement.

Ref. 1 - Noise Measurement and generation by Paul Wade (W1GHZ) previously published in the Microwave Newsletter but now available online (www.qsi.net/n1bwt/noise99.pdf).
Ref. 2 - Simple PANFI design – "an Alignment aid" in the Microwave Handbook, RSGB, 1993 reprint, Vol.2 page 10.37 to 10.42 . It should be noted that some modifications were made to this.

1. The Loudspeaker was omitted from the final build. The LS was found to be acting as a microphone and inputting sound into the system – this gave spurious readings. Therefore an 8 ohm resistor was used instead.
2. The 100uA meter was provided with 8 and 16dB scales and appropriate changes to the series resistors were made.
3. A mains powered +/-17V supply was used to give a regulated 15V for the Head and regulated +/- 12V for the rest of the circuit.

Low Cost Noise Source for VHF to 10GHz

Based on Qualcomm surplus material

Kerry Banke, N6IZW

During Our November 2000 San Diego Microwave Group meeting we held a session on noise figure measurement. During problem solving activities in the following weeks, it became obvious that others in our group could make good use of a calibrated noise source. I began looking for an alternative to loaning out the noise source used with my Sanders noise figure meter. I did a quick scan of articles on home brew noise generators and studied in particular Paul Wade's article "Noise: Measurement and Generation" which is available on his website as well as in the ARRL 2nd Microwave Projects Manual.

In his excellent article, Paul describes what he went through to build various homebrew noise generators, which included dismantling a commercial unit. As soon as I finished the article, an idea hit me that I felt might solve the problems of obtaining high frequency diodes and high quality microwave circuit board material as described in the article. My thought was to use a small section cut from a 14.5GHz upconverter mixer circuit board used in the early Qualcomm Omnitraks units. This mixer is not useful at 10GHz and is not readily modified to do so but does contain two very good microwave diodes (good through 24GHz) mounted on 15 mil Teflon circuit board

The idea was to see how well this mixer diode would perform as a broadband noise source when operated at the breakdown point with reverse bias. I used Paul's circuit which has the anode of the diode connected directly to the 20dB attenuator. The cathode is RF bypassed to ground and receives bias through a series resistor. Following Paul's article I found that for the first unit 1.2 milliamps produced the maximum output at 10 GHz so a series resistor was selected (1.6K) to provide 1.2ma with a 12v power supply.

Probably the most critical aspect of construction is having as short of paths to ground as possible for the RF bypassing and to the SMA output connector. The 0.5 pF capacitor needs to be a good quality microwave type and is removed from another part of the 14.5 GHz mixer board. I made the

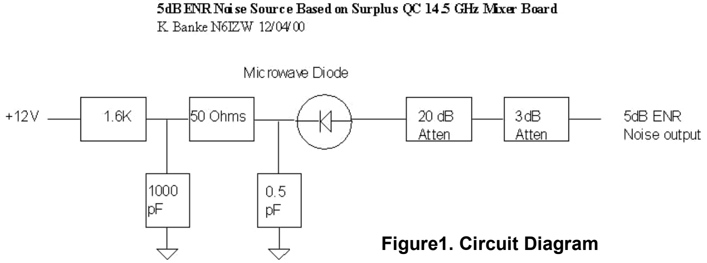

5dB ENR Noise Source Based on Surplus QC 14.5 GHz Mixer Board
K Banke N6IZW 12/04/00

Figure1. Circuit Diagram

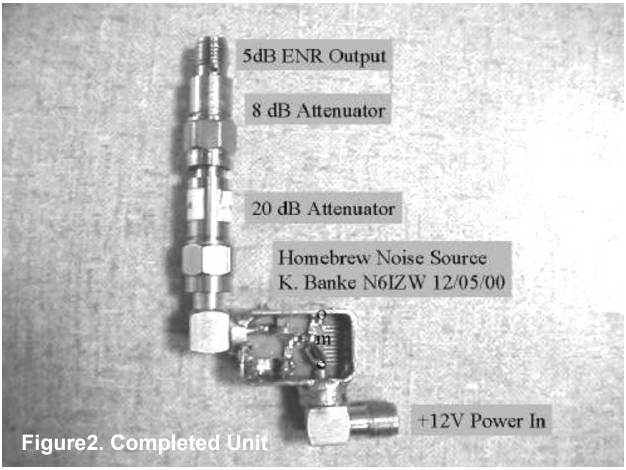


Figure2. Completed Unit

housing as small as possible to prevent unwanted resonances and also aid in good grounding. The quality of the output attenuator is also a big factor in flatness of frequency response.

Half of the battle is building a good noise source & the other half is calibrating it. The approach I decided to use for calibration over the 145 MHz - 10 GHz Ham bands was to use enough broad band amplification ahead of a spectrum analyser (set for 1 dB/div) to provide a Y-Factor of about 3 dB with a source ENR of 5dB (about a 5 dB NF). My setup consisted of two, three stage MMIC amps built by removing the FETs from QC surplus "GOLD Board" LNAs and replacing them with HP MGA-86576 devices. I found that working with two broadband amps with considerable gain creates intermodulation products from noise when used without a filter between the amplifiers. For 145 MHz-3456 MHz I was able to use one amplifier without a filter and controlled the gain by varying the power supply voltage. The gain was increased only as needed to get a Y-Factor in the order of at least 3dB. For 5.7 and 10 GHz, a filter was inserted between the two amplifiers and the gain again adjusted to the minimum capable of doing the job. A goal of 5dB ENR was set as this should be useful for measurement of most preamps in the 1- 10 dB NF range. It was found that a total of 23 dB of attenuation matched the output of the uncalibrated unit to the calibrated noise source at an arbitrary frequency of 2304 MHz. Other values could certainly be used to match the requirements of particular noise figure measuring units.

My experience with the Sanders unit is that it is quite fussy about levels and such and may give erroneous readings unless careful attention is given to the measurement setup. For this reason I tend to use a spectrum analyser or power meter to measure the Y-Factor and calculate the Noise figure. This is not very convenient though when trying to make adjustments on the fly but allows measurement of most amplifiers and transverters.

I generated an Excel spreadsheet (available from Martyn Kinder G0CZD) or
http://www.ham-radio.com/sbms/sd/nfcalc2.zip)
to aid in calibrating against a known

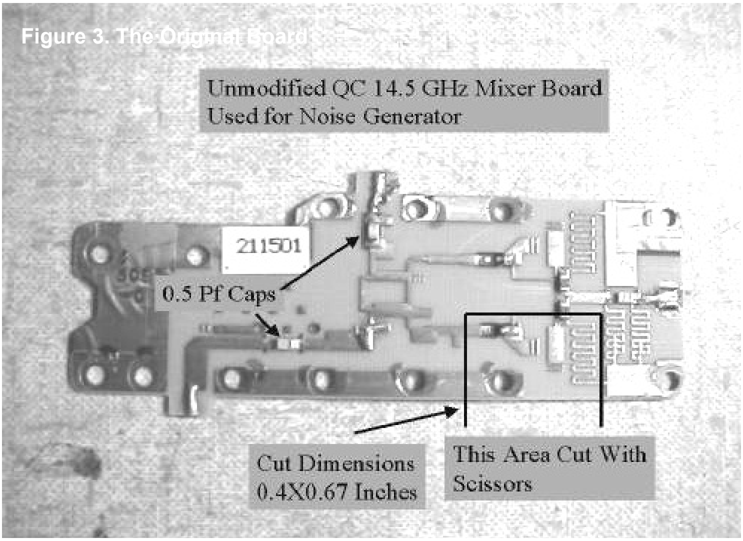


Figure 3. The Original Board

source. The Y-Factor was recorded for both a calibrated source (5 dB ENR) and the uncalibrated unit at each frequency of interest. The Y-Factor values along with the ENR of the calibrated source were entered into the sheet which calculated the ENR of the uncalibrated unit. Other approaches certainly could be used such as using a transverter with the noise sources and measuring the Y-factor at the IF or at the audio output of the IF radio. The results of the first units showed a variation of +/- 1.5 dB over the 50 MHz to 10 GHz range. I believe that with more work the output can be made even flatter across the range.

Construction:

There are actually two useable diodes on the mixer board. Only one is mounted as to easily accept a positive bias supply. The other can be used either with a negative supply or can be carefully removed and re-soldered in the opposite direction. The white dot (cathode) needs to be connected to the RF bypass/ bias network for a positive supply.

Remove the Chip capacitors from the mixer board before cutting with scissors. (See Figure 3.)

The board dimensions of 0.4X0.67 inches are not too critical but should be followed to keep the RF paths very short. After cutting the board, cut/ remove the two traces shown in the picture. Cut the centre pin off the output SMA connector so as to not overlap onto the diode. Position & solder the centre pin onto the board trace and then solder the outside of the connector to the bottom of the board. Use 0.5 inch wide hobby brass which is .01-. 025 thick to make a box around the board as shown. Position the brass & then tack in place in a couple of locations until satisfied with the positioning. Solder the brass to the bottom side of the board and also the output connector. Drill a 0.065 dia hole in about the position shown to mount the bias connector. Small coax could probably also be run directly out of the noise generator for bias rather than using a SMA connector. Install the capacitors and resistors as

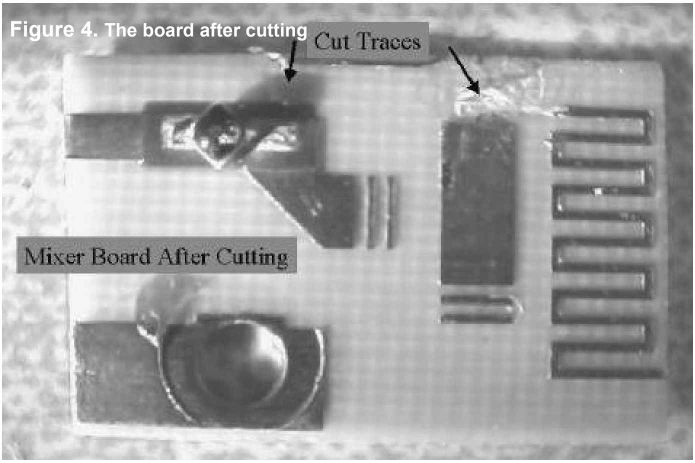


Figure 4. The board after cutting

shown in Figure 5. Temporarily install a 1K resistor in place of the 1.6 K resistor until the required bias current is determined. Connect a current meter in series with the 1-k resistor and a variable supply. Turn on the supply and increase the voltage until the current reaches about 1 milliamp.

Connect the noise generator output through about 25 dB of attenuation to the setup which will be used to calibrate the ENR and adjust the supply voltage for maximum noise output at the maximum frequency of interest (I used 10368 MHz). Replace the 1k resistor with a resistor as required to operate the noise source at that current form the desired bias supply (I used a regulated 12V).

Calibration

Connect the calibrated noise source (5dB ENR assumed at this point) to the test setup and note the Y-factor. Connect the uncalibrated noise source and change the 25-dB attenuator as required to match the Y- Factor of the calibrated source as close as possible. I performed calibration at the following frequencies which were of particular interest for me and for which I had equipment available: 50 MHz, 145 MHz, 436 MHz, 1296 MHz, 2304 MHz, 3456 MHz, 5760 MHz, 10368 MHz. Paul's article goes into the precautions that must be taken when using broad band amplifiers with noise sources to prevent erroneous readings. At each frequency of calibration, the Y-Factor is noted for both the calibrated and uncalibrated source and entered in to the spreadsheet to calculate the ENR for the uncalibrated unit.

The formulas used in the spreadsheet are given below:

Noise Figure Calculation

NF = ENR-10Log(Y-1)

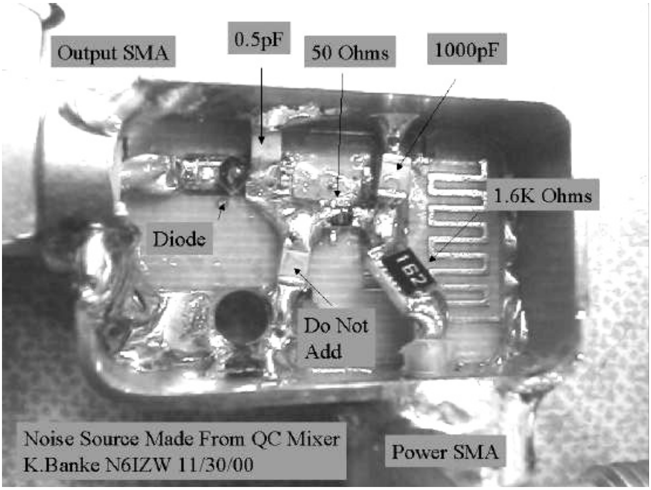
NF = Noise Figure in dB
ENR = Excess noise source in dB
Y = ratio of levels with noise source on/off (not dB)

Equation used for Excel program to calculate NF:

NF = ENR-(10*Log(10^(YdB/10)-1))

YdB is the ratio of levels with noise source on/off in dB

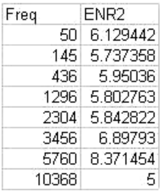
Figure 5. A close-up view of the assembled unit.



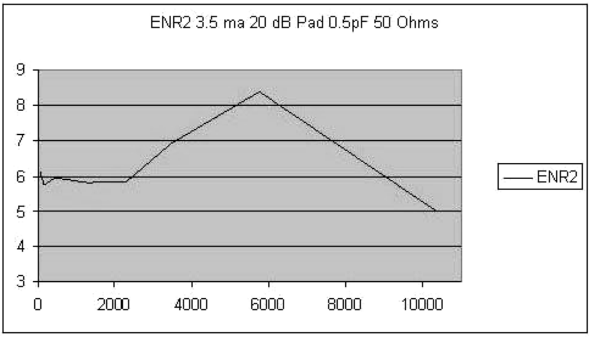
Equation used for calibrating an unknown ENR source against a calibrated unit.

ENRucs=(ENRcs-10*(LOG(10^(YdBcs/10)- 1))+ (10*(LOG(10^(YdBucs/10)- 1)))

ENRucs is the calculated ENR in dB of the uncalibrated source.
ENRcs is the ENR in dB of the calibrated source.
YdBcs is the Y-Factor (dB) produced by the calibrated source.
YdBucs is the Y-Factor (dB) produced by the uncalibrated source.



Freq	ENR2
50	6.129442
145	5.737358
436	5.95036
1296	5.802763
2304	5.842822
3456	6.89793
5760	8.371454
10368	5



A Noise Power Indicator

Brian Coleman. G4NNS

A noise power indicator can be a useful tool for judging the performance of microwave receivers using the Y factor method (cold sky vs. warm object). This relies on the fact that all bodies above absolute zero emit "black body" radiation which can be detected at the higher microwave frequencies.

In practice this means that you can use the "cold" sky vs. a known "warm" reference. I find aiming the antenna at the house nearby is best and you can assume its temperature is about 293K. Measurement as opposed to indication can be accomplished by using a precision attenuator in front of the indicator and using the same level at the detector to avoid any non-linearity in the system. I also use my system to ensure that my EME antenna is tracking correctly using moon noise of which I see about 2.5dB at 10GHz. Other uses of this tool can include antenna testing. It provides a more precise and easier to read system than an S-meter and can, in conjunction with a precision attenuator provide a means of making fairly precise measurements of the order of +/- 0.2dB. When using a warm source such as your house remember that it should fill the aperture of your antenna. This is why small antennas (more than about 0.5 degrees beam width) do not see the full potential sun or moon noise.

Circuit description (see Figure 1)

The system consists of a high gain amplifier (circa 70dBs) and a sensitive detector at the IF frequency (in this case 144MHz). The bandwidth needs to be as wide as possible but not exceeding that of the transverter front end. This is important as noise generated outside the desired pass band of the transverter is not relevant to the system performance and could swamp the power indication and lead to misleadingly poor indication of receiver performance. The 3dB bandwidth of this indicator is about 1MHz. Relative measurements are made by using a switched attenuator in front of the noise power indicator. Such attenuators can be found at microwave flea markets.

The main challenges to the circuit design and construction are:-

1) minimising the gain required to drive the detector. I.e. the detector must be as sensitive as possible and

2) keeping the gain stages stable at all frequencies including those outside the intended pass band.

The first requirement is met by using forward bias on the schottky detector diode to move it into its linear and most sensitive region and by using a simple op-amp to amplify the DC difference between the RF detector and another matched reference diode, similarly forward biased but decoupled to RF, in the same SM package.

The second requirement is achieved by using MMIC gain stages with careful construction, grounding and shielding. With so much gain on one frequency, great care has to be taken in the layout and construction to avoid instability. The stage before the detector and second band pass filter uses a JFET as this is more tolerant of impedance mismatch.

No originality is claimed for any part of the circuit but some time and effort has been spent in trying to ensure the repeatability of the layout and simplicity of the PCB. If good practice for RF construction is followed, it should be possible to build the circuit and get it to be stable quite easily. The input must however look substantially resistive and be correctly matched at all frequencies. This can be achieved by the use of some fixed attenuation at the input.

Construction (see Figure 2)

A simple way to make the PCB is to cover the double sided copper laminate with sticky tape such as 3M Magic tape then stick a print of the track layout onto the tape and use a scalpel or sharp modelling knife to cut out and peel off the areas to be exposed to the etchant.

The design uses as many straight lines as possible to simplify this. I found it helpful to use a straight edge under a 3D magnifier. After etching and cleaning the PCB, the screens should be cut to size. The long, side screens are best made using 1/16" laminate as used for the main PCB. The inter stage screens are best made from 1/32" copper laminate or from tin plate. I did not find end screens to be necessary but if you fit them they should be made from 1/16" laminate. Drill 2.5mm holes to recess the MAR8s, which should be the "drop in" type, to minimise lead lengths. Also drill 0.8mm holes for the Via pins. These pins are essential for stability and it is possible that more than the four shown may be needed. (see setting up).

As with any PCB construction it is best to start with the smallest components first as the larger ones will otherwise get in the way **(see figures 3 & 4)**. Whilst the screens should be fitted last you may find it easier to fit L1 and L2 after fitting the long side screen closest to these inductors as this will make it easier to solder the seam. L1 and L2 are separated by about 3mm and are 2-3mm above the ground plane. Note also that the long side screens extend below the main PCB and should match the height of the inter-stage screens on the component side. They should be soldered both to the top and bottom surfaces of the main PCB. Make sure the side screens remain at 90 degrees to the PCB by "tack soldering" both sides and repositioning as necessary before soldering the entire length of the seams.

When assembling the DC amplifier it is again best to fit the smallest components including the wire links and pins first. I find it best to make the track cuts after the components are in place and to use this process as a check for correct placement of the components. Start fitting the active components with the +5V regulator, and test this before fitting the voltage converter. Test that the + and – 10V rails are working before fitting the op-amp.

Setting up

Terminate the input with a good 50 Ohm load. This is essential for stability. Adjust VR1 for zero reading on the meter. If this is not possible the amplifier may be oscillating. This should not be happening! If it is try touching different gain stages to see if the meter reading changes. You may be able to identify which stage is oscillating by this method. You could try lifting one end of each one Ohm resistor in turn to identify the oscillating stage. Whatever you try, make sure it is reversible and restore it to the original condition before trying something else.

Here are some things to try if necessary:

1) Change the capacitor and inductor tuning to see if the oscillation stops.
2) Add additional via pins close to the MAR8s. Those indicated in the design have proved sufficient but in some cases more may be required to achieve stability.
3) Try adjusting the values for R1, R6 or R7 for the stage that is oscillating.

Once the amplifier is stable and the meter set to zero, apply a small signal within the pass-band e.g. 144.50MHz from a 50 Ohm source through a switched attenuator. Tune VC1, VC2, L3 and L4 for maximum reading increasing the attenuation as you go to keep the meter on scale. The prototypes, when set up as described, had a Full Scale reading on the most sensitive range, corresponding to about –80 to –85dBm.

Next, connect to the receiver or transverter under test. The Indicator should show a non zero reading when the transverter is switched on but make sure there are no strong signals in the pass-band, such as local beacons. If the meter is on the end stop try inserting some attenuation between the transverter and the indicator

Applications

Receiver performance can be measured using ground noise and quiet sky noise. For EME systems where the antenna beam width is close to or less than the angular diameter of the source, Sun and moon noise measurements provide a useful indication of receive performance. The indicator can also be used for calibration tracking systems. The indicator also has potential for plotting polar diagrams and measuring gain of antennas for 2m or bands where a transverter has a 2m IF.

For EME, the indicator can display moon noise during receive periods to provide confirmation that the tracking system is functioning correctly.

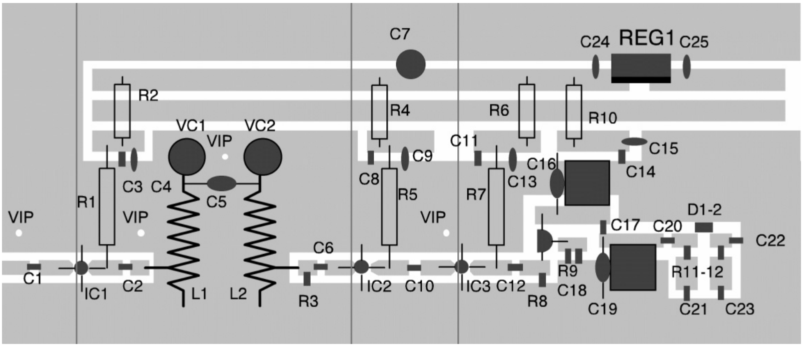
For this purpose, it is necessary to supply a sample of the IF to the indicator through a splitter (T) and to switch the indicator out of circuit during transmit periods. It is desirable to switch off some of gain stages during transmit periods, to avoid overload. Some gain drift may be experienced but this is minimised if the smallest number of stages are switched off. Try switching off just the first one or two stages if possible. A spare power track is supplied on the PCB for this purpose. The circuit diagram shows the possible arrangement for RF switching. The number of stages to be switched will depend on the isolation of the relay used.

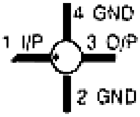
Notes

1. If you fit the unit in a box and it becomes unstable, try lining the box with conducting foam or lossy rubber where it is closest to the component side of the Noise amplifier.
2. On the supplied Veroboard component layout for the DC amplifier, **C26** is missing. This decoupling capacitor can be conveniently soldered across the MX680 IC, from pin 8 to pin 5.
3. Further information, if required can be obtained from Brian, G4NNS, QTHR or email to:

Brian-coleman@tiscali.co.uk

4. A component list, **figure 4**, is provided at the end of this article.
5. Except for the MAR8 modamps, semiconductor pin outs are shown on the circuit diagram.

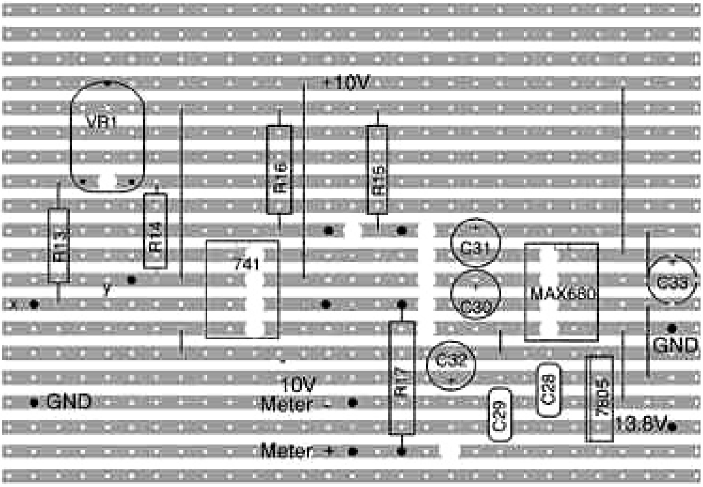


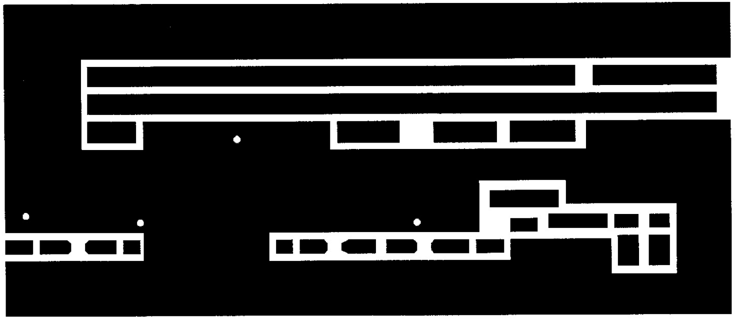


MAR8 pin out

Above:
Main amplifier pcb track and component layout

Below: DC amplifier Veroboard layout. Note that C26 is missing from this diagram and should be soldered from the +10V line to ground at a convenient point (e.g. across pins 5 and 8 of the MAX680)





Main PCB 1/16 Laminate 105 x 45mm
Note position of 4 via holes.

Intermediate Screens
1/32” Laminate
45 x 20mm
QTY 3

Side screens 1/16” Laminate 105 x 25mm . . . QTY 2

G4NNS NOISE AMPLIFIER
Component List

R1 100R Ax
R2 1R Ax
R3 51R SMD
R4 1R Ax
R5 120R Ax
R6 1R Ax
R7 82R Ax
R8 100K SM
R9 2K2 SM
R10 100R Ax
R11 10K SM
R12 10K SM
R13 1M0 Ax
R14 1M0 Ax
VR1 1M
R15 1M Ax
R16 100K Ax
R17 SOT

VR1 1M0

C1 1nF SM
C2 1nF SM
C3 1nF SM
C4 100nF SM
C5 3.3pF Rad
C6 1nF SM
C7 10uf Tant
C8 1nF SM
C9 100nF SM
C10 1nF SM
C11 1nF SM
C12 1nF SM
C13 100nF SM
C14 1nF SM
C15 100nF SM
C16 33pf Rad
C17 2.2pF SM
C18 1nF SM
C19 33pF Rad
C20 22pF SM

C21 1nF SM

VC122pF
VC2 22pF

L1 6T 6mm id
L2 6T 6mm id
L3 MC119 1.5T
L4 MC119 1.5T

Q1 J310
IC4 741
IC5 MAX 680 Farnell 246_451
REG1 7810
REG2 7805
IC1 MAR8
IC2 MAR8
IC3 MAR8
D1-2 BAT 54C Farnell 302_284

C22 1nF SM
C23 1nF SM
C24 100nF SM
C25 100nF SM
C26 100nF Rad
C27 100nF Rad
C28 100nF Rad
C29 100nF Rad
C30 4.7 uF
C31 4.7 uF
C32 4.7 uF
C33 4.7 uF

Some Observations on Sun Noise Measurements

Neil Whiting, G4BRK

Introduction

Sun noise is an accepted way to gauge system performance on microwave frequencies, but apart from EME station reports there are few indications of what is a reasonable result on typical amateur microwave systems. In the course of checking and optimising feed positions on a selection of dishes I made a number of measurements, which may be of interest to others.

Setup

Several home-built transverters were used to convert the various microwave bands down to 144MHz. My RATS system (G4PMK design) was used as a sensitive level meter on the transverter 144MHz output – the optional MAR-6 amplifier was required between transverter output and RATS receiver in order to get sufficient signal. A calibrated attenuator (0-11 dB in 0.1dB steps) after the transverter served to give an accurate measurement of the Y-factor (difference between cold-sky and sun noise).

Measurements

The measurements were made by aiming the dish away from the sun and setting the RATS offset to get a convenient, low scale reading (cold sky) with the attenuator set to 0dB. The dish was then pointed at the sun and the reading maximised. The attenuator

Antenna	Band	RXNF at ant dB	Antenna gain dBi	Y factor dB	Theoretical YFactor dB
1.2m prime focus	2320	2.2	27.9	2.0	3.9
0.5f/d, RMX feed	5760	3.5	35.8	0.8	2.5
90cm prime focus	2320	2.2	25.4	2.0	2.6
0.4f/d, RMX feed	5760	3.5	33.3	0.5	1.6
90cm prime focus	2320	2.2	25.4	1.5	2.6
0.6f/d, RMX feed	5760	3.5	33.3	0.5	1.6
80cm offset,	10368	1.5	37.1	0.5	2.2
10/24G feed	24192	3	44.5	2.0	0.8
80cm offset,	2320	1.2	24.1	2.5	2.1
RMX feed				0.5	
60cm offset,	10368		34.8	3.5	1.8
Amstrad feed					
35el Tonna Yagi	1296	1.2	17	1.5	2.6
	1296	1.2	17	2.0	2.6
67el Wimo Yagi	1296	1.2	22	1.6	5.3
	1296	1.2	22	2.6	5.3

was then adjusted to give the same reading as the cold sky. The attenuation used gives the Y-factor.

Antenna gains for the dishes and theoretical Y-Factor come from the Ed Cole (AL7EB) spreadsheet available on the WWW. Dish efficiency of 55% is assumed. For the estimates of 23cm Yagi gains, the Tonna is based on simulation and the Wimo on manufacturers claims.

RX noise figures are estimates.

Conclusions

I set out to confirm feed positions for my dishes, and this was successful. It was noted that the deeper dish was more critical on feed position – 10% movement in distance from the dish is quite noticeable in the measurements, though only fractions of a dB. This contrasts with results by others which suggest position needs to be correct to a few mm. I needed 4cm on the 1.2m dish before I could see the Y-factor drop by 0.1dB. The measurements also raise a number of questions, some of which I can attempt to answer:
On 2320MHz – why is the 90cm,0.4f/d dish as good as the 1.2m?

The WA3RMX feed is designed for deeper dishes, 0.3-0.4f/d, so over-illuminates the 1.2m dish. This adds ground noise from behind the dish to the cold-sky reading and less increase is therefore seen when aiming at the sun.

On 5760MHz, why is the 1.2m dish showing an improvement?
In this case I can only guess that the RMX feed gives a narrower pattern on 5760 and is relatively better on the shallower dishes.

Why is the 60cm dish better than the 80cm dish on 10368MHz?
The Amstrad feed on the 60cm is obviously working very well. The 10/24GHz feed on the 80cm uses a piece of 24GHz waveguide let into the back wall of a G4DDK feed. The 'DDK feed is designed for ~0.5f/d, so will be over-illuminating the 80cm dish, reducing the Y-factor due to ground noise contribution. There may also be a reduction is efficiency due to the 24GHz guide. The 24GHz figure, on the other hand, seems very good – it looks like this feed is accidentally working quite well!

How come the 67el Yagi is not much better than the 35el? I don't know. The difference in claimed gains is ~3dB and the 35el simulates as worse than claimed. There may be a matching difference causing a change in the effective noise figure – I seem to get a higher background noise on the 67el. Some more investigation is required.

It is interesting that I see more Y-factor when the antennas are mounted horizontally and moved towards the setting sun. I would have expected less due to part of the beam including the ground and increasing the off-sun background noise. It may be that I am seeing the effect of ground-gain – something that is seen by 2m EME stations. Any other ideas? I learned something from making these measurements and can use them in the future to check the effect of improvements in each of the systems.

Chapter 14

Surplus

Surplus

Building Blocks

Compact Power Supply for the Qualcomm Omnitrak Amplifier

Dave Robinson WW2R

I thought it was time to redo my 10GHz portable unit to increase power without increasing the size, so that it could continue to be mounted at mast-head. It was decided to use the Qualcomm Omnitrak PA (ref 1) with the DB6NT original transverter design (ref 2), G4DDK004 oscillator and the G3WDG preamp (ref 3).

There was insufficient space in the box for the Omnitrak PSU board that came with the amp, especially as its TX/RX switching was already provided in the DB6NT module. I therefore needed to devise a small PCB, which could provide 10V at 1A max and –5V at around 50mA.

For the positive rail, the LT1086 1.5A low volt drop regulator would be used again (ref 4). However the usual 7660 chip was not capable. Looking through some regulators data sheets (ref 5) I came across the LT1054 chip which had many advantages over its predecessors: -

1 .It is capable of providing 100mA (7660 max is around 20mA)

2. It is rated for an input voltage of 15V maximum so can be run directly off 12V supply (7660 maximum is 10V)

3. Most importantly, it has an internal regulator that regulates the negative output voltage. (You don’t need to worry about the 120 ohm internal resistance of 7660)

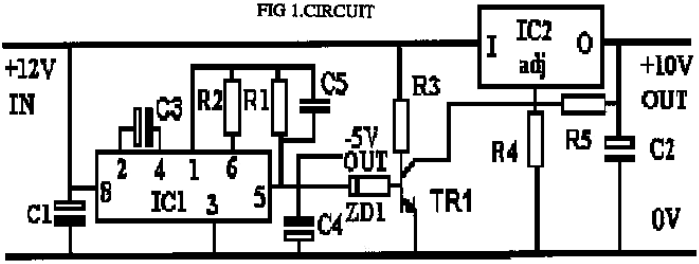
Accordingly, the circuit shown in **Fig 1** was devised. Failure of the negative supply reduces the positive supply to around 1.2V. The component listing is shown in **Table 1**. The PCB layout is shown in **Fig 2** and the overlay in **Fig 3**. Note that the components are mounted on the track side of the board and that pin 7 of IC3 is chopped off and not soldered to the track below it. Surface mount resistors and capacitors can be used on the board. The PCB was mounted on the back of the amplifier casting (the back of which had been milled flat) taking care to insulate the metal tab of IC2 from ground.
The output voltages of the 3 modules built so far were measured as 9.82 and –4.89.

Purists may be tempted to adjust the resistor values slightly to achieve exactly 10.0V and –5.0V. I tried it, but doing so made no difference at all to the Qualcom’s 1.1W output power, probably due to its on board bias regulators.

References:

1. http://www.ntplx.net/~wz1v/w1ril.html
2. Dubus Technik III pp324-333
3. Dubus Technik IV pp276-339 (also Proceedings Microwave update 1991 and 1995)
4. "Using the California Microwave 11-026700 transmitter assembly" Microwave
update proceedings 1995
5. http://www.linear.com/prodinfo/dslist.html

FIG 1.CIRCUIT



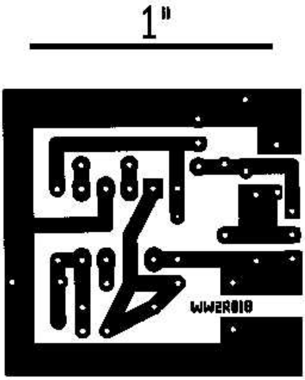


FIG 2. PCB LAYOUT

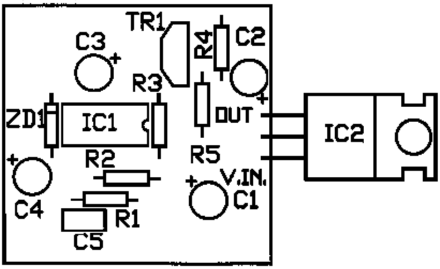


FIG 3. COMPONENT OVERLAY

Table 1: Component list

Resistors	**Caps**	**Semiconductors**
R1 100k	C1 10uF 25V	IC1 LT1054
R2 20k	C2 10uF 25V	IC2 LT1086
R3 10k	C3 10uF 25V	ZD1 4.7V 0.4W
R4 820R	C4 100uF 10V tantalum	TR1 2N2222a
R5 120R	C5 2200pF	

Sub 2dB Noise Figure for the Drake 2880 Downconverter

David Bowman, G0MRF

A satellite band Drake 2880, fitted with Toko filter, MGA86576 and 22pF caps in the I.F, 1.94dB noise figure and 38dB conversion gain…. interested ? - Then read on…

There have been a series of modifications detailed for the Drake 2880. The following improvement was tested at the AMSAT Colloquium held at the University of Surrey in July '99. This procedure is only suitable for Drakes operating in the Satellite sub band of 2400 or 2450 MHz. A second Drake modified was within 0.01dB ! This modification is consistent.

It is assumed that the following standard changes have been carried out:

1. Crystal changed and IF coils and chip caps removed - JN1GKZ
2. 2 x 22pF added to IF - G0MRF
3. Front end FET replaced with Hewlett Packard MGA86576.

After these modifications you should already have improved the performance to beyond that of the original manufacturer's specification. 2.4dB noise figure with a 28dB conversion gain was the average measured over a two day period.

Assessing the circuit for a next step, I identified the image filter as a source of excessive signal loss. The 86576 under optimum conditions should be able to produce a NF of 1.6dB. The Drake's image filter has around 4dB of insertion loss and this is undesirable when the front end is a single device. Additionally, we also saw some interaction between the stripline filter and the small cover which isolates the RF / IF parts of the circuit. In some cases we measured up to 2dB difference in noise figure between "lid on and lid off". So the stripline filter had to go!

Finding a replacement wasn't too difficult. A couple of years back I designed a 2.4 GHz to 432MHz converter which used a Toko dielectric filter. The filter is centred on 2450 MHz and is used in wireless local area networks (LANs) It has a pass band of +/- 50MHz but more importantly, the insertion loss is only 1dB.

To change the filter requires a little surgery with a modelling knife, the removal of the 3 striplines, fitting 2 ground pins and the repositioning of the mixer coupling capacitor.

Procedure

1. Desolder mixer coupling capacitor which is close to the output stripline.- Keep for later!
2. Cut across the input stripline with a sharp knife. This extends the 50 Ohm track which feeds the filter by about 3mm (The width of the stripline)
3. Cut the track between the output stripline and the mixer in the position shown in the photo (see next page).
4. Remove unwanted parts of the striplines by heating with a soldering iron and lifting the copper clear of the board.
5. Position the filter on the board and carefully mark the positions for the two ground pins.
6. Drill the PCB with a 1mm drill ensuring that the drill also marks the die cast case under the board.
7. Remove the board and fit / solder the pins from the top. Cut flush on the underside of the board.
8. Take a larger drill, 4 or 5 mm, and look for the marks in the casing. re-

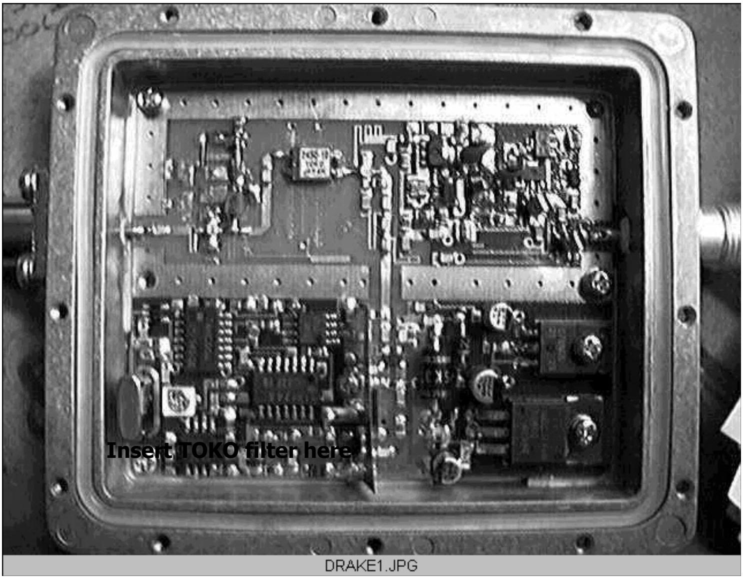
move a small amount of metal from the casing to allow any protrusion from the pins to be recessed into the case. - This allows the board to lay flat when refitted.

9. Refit board into case. Solder filter to 1 track and the heads of the two pins.

10. Solder a coupling capacitor directly to the filter output terminal and to the track feeding the mixer. Replace all covers / screws.

You now have one of the hottest Drakes around!

THE DRAKE 2880 DOWNCONVERTER



DRAKE1.JPG

A Checklist for putting the Toshiba 20 and 40 watt 3.4GHz Linear Amplifiers on the Air

John Jaminet, W3HMS

Introduction

This paper was created from several messages received on the Internet Lists in September/ October 2001 by Steve, **N2CEI**, at DEMI, Rick, **K1DS** and an article in the MUD 2003 Proceedings, pages 66-70, by Dave, **WW2R**. Important new info was received from Owen, **K6LEW** in November 2001 on the importance of voltage regulation. It is incorporated in the text. In addition, our own **N3TWT-ATV** experiences with 24/7 operation in FM ATV (key down for months on end!) service are reflected in this update.

Basic Information

There are **two** versions of this amplifier:
a. 20 Watts output @ 3.4GHz with 1mW input.
b. 40 Watts output @ 3.4GHz with 1mW input.
Supply Voltage: + 12.6VDC (exactly)

Vendor Info on the 20 Watt Amplifier

This is a Toshiba 20W Linear Microwave Amplifier for use in the 3.44 to 3.68 GHz range. It is sold in the original manufacturers packaging.

Heat sinking: Required…it gets hot..
Size: 5" x 8" x 1 ".
Weight: About 2 pounds.
Input power: For full output, this is about 1 milliwatt or 0dBm.
Specifications: The specs on these amps are very "tight" and are typically as follows:
Linear Gain = 42.5dB; Gain: Ripple = .1dB
Linearity is -45dB; Return Loss (in and out)= 25dB
DC Power Supply current: Typically +12 VDC at 10.6 amps.
Power Jack: The DC power and control connector is a DB-15 male.
DB-15 Pinout: There are 3 rows of 5 pins each.

The ground is made from pins 3,7,10,11 connected together.
The +12VDC lead is made from pins 1,2,12,13 connected together All VCC and ground pins need to be connected to handle the 10.6 amps.
Pin 9 is the enable pin (TTL) which must be connected to ground in transmit to switch the internal power supply on.
When pin 9 is not grounded, the 12VDC supply draws about 15mA. and the amp is in stand-by mode.
This is a Class A amp and as a linear amp, it will draw about 10.6A with no signal input. Pins 4,5,8,15 are assorted alarm output pins low true….no more is known.

The 40 Watt Amplifier

This is a new Toshiba UM2683A 40W Linear Microwave Amplifier for use in the 3.4 to 3.6GHz range. It is sold in the original manufacturers packaging. This amplifier differs from the "2683B" "20W" version on other auctions because there is a TMD0305-2 MMIC instead of the discrete circuitry in the front end. The TMD305-2 part is a 3.4-5.1GHz amp with 2 watt output and 22dB Power Gain. Turning the two pots at the far left on the lower board in the photo fully clockwise (shutting down the attenuator) and peaking the power with the third pot (2nd and 3rd stage bias) yielded 46dBm, 40 watts @ 3.456GHz using 12.0-12.6VDC (as measured at the connector) with about 0dBm input power.

Power supply requirement is 12.6VDC @ 15amps after readjusting gain and bias.

Size

The size of these amps is 5 x 8 x 1 inch and weight is almost 2 pounds. Input power required for full output is about 0dBm. DC power and control connector is a DB-15 male.

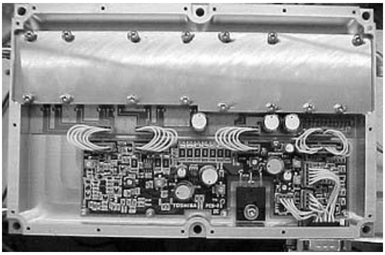
Pinout is as follows:

Ground are pins 3,7,10,11; +12VDC are pins 1,2,12,13;
All VCC and ground pins need to be connected to handle the 15 amps.
Pin 9 is the enable pin (TTL) which must be grounded to switch the internal supply on.

When not grounded, the 12VDC supply draws only about 15mA. and the amp is in a stand-by mode. Since this is a linear amp, it will draw about 14-15 amps without signal input. Pins 4,5,8,15 are assorted alarm output pins low true…no more is known. Only new amps are being shipped.

Heat Sink Requirements

These amplifiers get VERY HOT! Heat sinking is required, particularly for FM ATV service many hours per day. We have found that the heat sink should be about 2.5 or more times the size of the amp and be blown with two high cfm blowers of no less than about 85 cfm capacity. We use two 12VDC blowers for twice the air flow and for redundancy in case of the failure of one blower.



In physics, we learned that heat rises and so it does indeed! Resist the temptation to mount the fins anyway but pointing UP as we have seen some disastrous results with the fins pointing down.

Even with fins up and 2 blowers we find the amp runs about 20 degrees higher than the temperature on the amp chassis, even in winter. We have mounted a small RS dual indoor and outdoor thermometer permanently on the chassis. We use the outdoor probe on the hottest part of the amp and heatsink. As I write this, an amp under test is showing 75F from the chassis probe and 55 F on the chassis. In summer, in a non-air conditioned space, the difference can be 30 degrees F. The results of poor heat removal have been small fires at the interior of the final amp shielding as revealed by the charred parts of two amps. Photos are available.

We recommend you use a very stiff power supply designed for 24 hour per day operation with at least twice the normal load of 15 amps. In our case this is 30 amps adjusted to give 12.6VDC on the amp terminals. Note that some commercial power supplies arrive giving 13.6VDC output. In some cases, the output is adjustable internally by a pot that is tough to find and adjust; a design indeed by Murphy, de Sade, or the village idiot.

Overall System View

When the amplifier is integrated into a

system such that relays, etc. are involved, the TX mode requires that the amplifier Pin 9, the PTT, be grounded for the amplifier to amplify, i.e. it turns on the power supply built into the amplifier. For RX, removing ground from Pin 9 places the amplifier into standby, drawing perhaps as much as 15mA, while the T/R relays all reposition for receive.

There is no feed-through via the amp. Changing from RX to TX involves just grounding Pin 9.

There are no relays internal to the amplifier for switching RF between RX and TX, hence, external relays for switching between RX and TX are required.

Critical Voltage Levels

Please refer to Step 9 (below) which is VERY critical. These amplifiers, especially the 40W version, will go into "fold back" and be hard to recognize in so doing. The 12.6 VDC +/- 0.2 is critical because the output FET regulator has an extremely narrow window due to the heavy current drain.

If you allow your primary voltage to get outside of this window, the FET bias voltage , as adjusted by R150 as measured on Pin 1 of the regulator interconnect to the amp section, you will note the need for drive in excess of 0dBm for 40dBm output (40W). (W3HMS used about 950 microwatts drive for 40 watts output). Excess drive is not good and indicates fold back which generates excess heat causing the "final" to generate more heat than it should.

These amps should draw 15 amps key down, not 14 amps, though some do vary. The set-up is simple enough but it has been necessary for several amps to be readjusted. This is because some operators did not see the importance of the relationship between the primary voltage at 12.6VDC

In summary

This is to say that R150 adjusts for 10.3VDC ONLY when the primary voltage is 12.6VDC with the amp under load (Pin 9 grounded) and the amp is drawing current.

If 12.6VDC in either direction is exceeded by about 0.3 - 0.4 VDC, adjusting R 150 will NOT regain 10.3VDC.

Thus, make absolutely certain you have sufficient heat sinking mounted to the amp for any tests or operation. The devices in this amp are extremely expensive and will not accept much heat. This amp gets VERY hot very quickly without a heat sink attached. One could also use a muffin fan blowing across the heat sink. One of the "alarm" functions available on the subminiature DB 15 appears to provide a "temperature" alarm. It is not known what the temperature should be when this function is energized but there is a voltage on one of the alarm lines that appears to relate to an increase in temperature. This amp does not self protect, i.e. shut down, so be very careful !

Perhaps the voltage could be used to control a relay which would remove PTT-ON from pin 9 but we do not know how much current it can handle. It may not be enough to control a relay, but it might control a NPNB transistor which could in turn control a relay, but this is not sure.

Output Power Monitoring

In an NTMS Feed point "article, also cited in MUD Proceedings 2003 on Page 66, Dave Robinson, WW2R, found a terminal in the amp which provides 5VDC equal to 40 watts output. This terminal is in a 6-pin connector at the right end of the amp, 3rd pin from left, when the shielded amplifier box is above the connector.

Another way to say it is the output connector will be on your right and the input connector on your left. This is the forward voltage, + voltage where 5VDC = 40 watts output. Doug connected a wire to an empty pin on the DB15 connector. I chose to drill a hole on the closest wall in a clear spot for a feedthrough capacitor and then connected a 0-10VDC meter via a shielded cable.

Conversion Steps

♦ Obtain power plug DB-15 male at RS or other. RS # is 276-1502 for about $2

♦ Find and install heat sink size about 5" by 8" (size of amp) or larger and muffin fan if desired

♦ Calculate and obtain the attenuator value needed if you have more than 1mW output from your rig. FYI, the DB6NT units have about 200+ mW output and I will use 24dB of attenuation. ♦ Connect DB 15 power plug pins # 3,7,10,11 to ground.

♦ Connect DB 15 power plug pins 1,2,12,13 as the + 12.6VDC lead usable at 15 amps load.

♦ Connect a switch or relay to ground pin #9 for XMT ♦ Install antenna relay and necessary DC power

♦ Remove cover to see the 4 pots R217, R210, R136 on 20 watt model, R138 on the 40 watt model, and R150. These first 3 are the first 3 counting from the left and R150 is directly to the right of the voltage regulator

♦ Counting from left, turn pots R217 and R 210 fully clockwise

♦ Connect exactly + 12.6VDC power and power up

♦ Measure +12.6VDC at pin 4 on P6 to ground

♦ Connect input signal on 3.4561GHz (3.4001 in Europe) at the level of 1mW or 0dBm

♦ Monitor supply current and regulate to 15.0 amps with R138 on 40W model and R136 on 20W model

♦ Connect up the antenna or dummy load using good quality N or SMA fittings*

♦ If needed, adjust R150 for 10. VDC (CRITICAL) on right hand pins of P3 and the 10.3VDC from the regulator controlled by R150 and measured at Pin 1. This 12.6VDC should be measured inside the unit with the cover off while the amp is under load, i.e. key down or Pin 9 grounded on the sub miniature DB 15.

Given that this pin is hard to use when the amp is fully set up, we found that we had about 0.09 VDC drop under load between the terminal strip on our chassis and internal pin 1 of Plug 6. Knowing this, we now do set up by putting 12.66VDC under load on our terminal strip. With the amplifier off the air, thus no load, we see 13.05VDC on our terminal strip.

NOTE: You do NOT need any RF drive for this set up as the amp runs in Class A. If you have a power supply with remote sensing use it. If it does either voltage or current drain sensing, use voltage sensing. The voltage MUST stay steady at 12.6VDC or something will surely be damaged, most probably the regulator, to be followed shortly thereafter by the costly output FET.

It is not necessary to monitor the drive level if you have reliable and repeatable equipment for the "exciter". Once you establish 0 dBm at the input port to the amplifier it should stay adjusted. If somehow the equipment is not reliable as to its output, and there is time to adjust, then monitoring the input level would be necessary.

Most modern transmitting equipment of today is reliable such that once power levels are set, the input level to the amp should remain set.

In an NTMS Feed point "article, also cited in MUD Proceedings 2003 on Page 66, Dave Robinson, WW2R, discussed his experience with this amp. In one tune up case of a new amp, we could not find the voltage controlled by

R150 as addressed above so we followed Dave’s advice to tune R150 and R138 for maximum output.

One could easily think about installing two VDC measurement jacks and make R150 an external pot so as to monitor and adjust both the 12.6 and 10.3VDC (at the VR) levels. No, the better choice is to monitor the output from a known and acceptable starting level. If you see the output begin to vary, then it is time to check the input level.

To clarify this point, R150 adjusts the output of the regulator BUT it does so in a very narrow window based on the primary input voltage of 12.6VDC and the amount of current being drawn and R138 will adjust this on the 40W version. ♦ Peak the power output with the third pot (This is R136 or R138) to yield 40 watts output

Testing

We have found that the output connectors can become quite HOT if the SWR is higher than desired, often too hot to touch! N fittings are desired but SMA fittings may simply have to be used per your individual needs. We recommend a careful “touchy/feely” exam even as little as 3 minutes after the on the air test commences.

Using the California Microwave 11-026700 transmitter assembly

Dave Robinson WW2R, G4FRE

Introduction

A considerable number of the California Microwave "Transmitter assembly 11-026700-08(h)" have been seen at rallies over the past 5 years. Many people have bought them, but it is hard to believe this, judging by the scarcity of "on-air" models. This paper details how to get them going on 3456MHz along with details of one method of incorporating it into a transverter and will hopefully encourage some to dust them off and help populate the band.

Equipment Description

In commercial use by AT&T for 3.7 - 4.2GHz microwave links, the amps were powered from positive ground 24Volts via a rack mounted California Microwave "Power supply 52-090095-0". The Power Amplifier unit consists of a 12.5 x 5.5 x 1.5 " aluminium can, containing the electronics, attached to a 13 x 6.75 x 1.5 " heatsink. The input connector is SMA, the output is via waveguide, but this is easily converted to SMA. A 6 core cable terminating in a 6 pin plastic plug supplies the operating voltages from the PSU.

The Power Supply unit consists of a 19 x 6.5 x2.5" box with one 2 pin connector, for applying -24V and a 6 pin connector for outputting the regulated voltages to the PA. It provides outputs of +10.25 and -5.3 Volts. It has an on/off switch and 4 test points.

The PSU is bigger than the PA, in addition it's -24 volt requirement makes using it portable a problem.

The first problem to be tackled therefore is to build a PSU to build the amp off 12V. In its original state the amplifier's performance is poor at 3456MHz and will need retuning.

Power Supply Unit

If the 6 screws securing the aluminium can of the PA are removed, along with the 6 smaller screws securing the aluminium screen below, it will be seen that the 6 core cable is connected as follows:

Orange and Purple	Drain Supply
Brown and Blue	Gate Bias
Grey	Ground
Red	rectified RF Sample

The orange and purple wires supply +10.25V at 1.7 and 1.5A respectively from isolated sources in the PSU. The 10.25V requirement precludes the use of the 78H series of regulators when running the PA off a lead acid battery as they require a minimum of a 2.5 Volt drop across them. A better idea is to use one of the LT108X series of low dropout voltage regulators which typically require 0.5 to 1V across them to maintain regulation. A single LT1084 regulator, which has a short circuit current of 5A is used.

The brown and blue wires supplying the -5.3V need around 10mA with the 10.25V supply on , 20mA with no 10.25V applied. This can be supplied by a LT1044 IC. The DC-DC converter chip is supplied from a 7808 regulator. The more usual ICL7660 is not used as it has been found more prone to expiring. The output of the LT1044 feeds an LM337T adjustable negative voltage regulator set to give -5.3V. The resistors around the LM337 have been increased, without adverse effect, from the usual recommended values, to reduce the current loading on the LT1044 IC.

It was decided to include extra circuitry to protect the amplifier from the loss of the negative rail, which has been known to destroy the amplifier. The circuit used COMPLETELY removes the positive 10.25V supply in the absence of the gate bias, rather than the more often published design such as (1) and (2), which only shuts down the regulator to 1.2V output.

The complete circuit diagram is shown in Fig 1. A PCB was designed for the circuit, excluding the components for the 10.25V regulators which are mounted directly onto the heatsinks, and the power relay. The PCB layout Fig 2. and the component overlay in Fig 3. The component listing is shown in Table 1.

Correct operation of the PSU should be ensured before proceeding. Normally the relay should operate and +10.25V and -5.3V should be measured at the appropriate points. If the -5.3V rail is removed (for example by disconnecting pin 8 of IC2) the relay should drop out and +10.5V rail should disappear.

Power Amplifier

Firstly the output connector has to be changed, to allow the usual RF connectors to be used on the amplifier. Carefully unsolder the brass pin from the output track of the amplifier. With a hex wrench, undo the 6 set screws holding the waveguide assembly in place and the four screws holding the tube to the edge of the PCB. It is possible, with a careful saw cut to leave part of the waveguide assembly with the 5 feed throughs and the earth lug and 4 mounting holes. For the faint hearted the safer option is to unsolder the 6 wires connected to the bracket and remove the whole assembly. This is replaced with a short length of 0.5" aluminium angle fitted with 4 bolt in feed throughs and a ground lug. (The brown and purple wires are connected to a common feed through). This method also shrinks the amplifiers area, allowing a changeover relay to be connected directly to the output connector.

The mounting plate left on the edge of the PCB has the correct fixing centres for a 4 hole SMA chassis mount female socket of the type with the long centre pin and the extended ptfe insulation. The centre pin and insulation are cut to length. The two holes for the lower screws are tapped and four long screws and 2 nuts hold the connector in place.

As a guide to the correct operation of the onboard PA regulator board measure the gate and Drain voltages of the devices:-

Device	Vgs	Vds
TR1	-1.1V	7.3V
TR2	-1.1V	8.1V
TR3	-1.2V	9.8V
TR4/5	-1.5V	9.8V

These voltages, are approximate and vary slightly between amplifiers, but give good indications of potential problems, and are measured with a 50 ohm load connected to the input and output of the amplifier to avoid instability corrupting the results.

Connect 10W power meter rated at least to 4GHz to the PA output. If a 3.9GHz source is available (such as a Midwest microwave "brick" awaiting conversion to 3456) lower the output to 1mW using attenuators as required (beware, some models produce over 250mW!) and apply to the PA input. The PA output should be at least 7.5W The next step is to apply a 1mW of 3456MHz to the amplifier. The output could be as low as 3W, so retuning is necessary. It will be observed that the board has extra printed stubs on the board, mostly on the inputs and outputs of the devices. These should be connected and disconnected with solder "blobs"

AFTER DISCONNECTING THE POWER SUPPLY BEFORE EACH ADJUSTMENT to maximize the output power working in sequence TR1-5, optimizing the gate matching first then moving onto the Drain. If necessary extra pieces of copper foil, available from hobby shops may be added to the board.

A recently discovered alternative to soldering the foil is to use adhesive backed copper foil which is available in various widths from stained glass craft shops. This was discovered to be the vital ingredient in my wife, Meg's N2NQI's newly mastered hobby and has meant many trips to supervise the correct widths being purchased. When properly tuned the amp should give at least 7W, it has been consistently found to give slightly less output at the lower frequency, despite repeated optimization efforts. Finally the inner cover should be replaced over the circuit boards, this has RF absorbing foam under it to maintain amplifier stability. The outer "can" was discarded.

The so far unused red wire drives the negative terminal of a 10mA meter via a series resistor to ground to indicate relative output.

Conclusions

Hopefully this article will encourage more activity on 3456 over longer distances. Remember the old adage of the ham bands "Use them or Lose them".

References

1. Power Amplifier for 13cm. E.Gobel, DUBUS 2/94 pp22-29

2.5GHz TV converters for amateur satellite reception

Sam Jewell, G4DDK

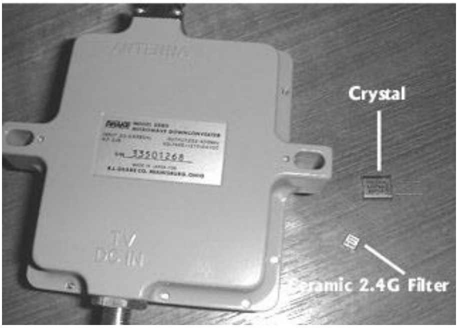


Figure 1.

Drake 2880 converter together with replacement crystal for the PLL and the ceramic 2.4GHz RF filter

Receive converters

Back in 1997 I published a construction article for a 2.3/2.4GHz receive converter for satellite and terrestrial use. The construction article was published in the Proceedings of the Weinheimer Tagung 42. This converter was similar to the design published in the RSGB Microwave Newsletter that same year by Dave Bowman, G0MRF, but I elected to use a 144MHz IF rather than 435MHz. I felt that a 144MHz IF was more appropriate for a mode S satellite receiver since it was my belief that 435MHz would be a more popular uplink band than 1269MHz. Consequently using 435MHz as the receiver IF would require the use of two 435MHz-capable transceivers (or transceiver and one multi-mode receiver/ scanner) in order to operate effectively. Time will tell if this assumption is right. Several months after I published my converter design I was at Microwave Update '98 in Sandusky, Ohio, when Toshi, JA1AAH and his friends appeared with several boxes of the Drake 2880 2.5GHz converters for sale at $25 each. Of course I had to have one.

When I got home I modified mine in accordance with the instructions on the JN1GKZ Website. The PLL crystal was changed to 8.8125MHz to give a local oscillator injection frequency of 2256MHz. The IF filter modifications of removing the two coils and capacitors was also done. I wasn't too impressed with the measured performance, but it was clear that with its compact weatherproof construction, it would be a very useful 2.4GHz converter when P3D finally got aloft. I put the Drake converter safely away, as the P3D launch was on hold yet again and I didn't have any immediate need for another 2.4GHz converter.

My own one worked just fine. Together with Lehane, G8KMH, I was providing the microwave measurements at the AMSAT-UK Satellite Colloquium at the University of Surrey in 1998. We were both surprised at the large number of Drake converters that appeared courtesy of Richard, G3RWL. Richard had already converted several units and Lehane and I, together with Dave, G0MRF, were able to measure their noise figure and insertion gain using a HP8970A noise figure meter and HP346A noise head. I think it would be true to say that the results were also a little disappointing. We measured converters both in unmodified form as well as those that had been modified according to the JN1GKZ instructions. The general spread of results showed that using the original published modifications, you could expect a noise figure of around 6.5dB and an insertion gain of about 18dB. These are highly averaged figures.

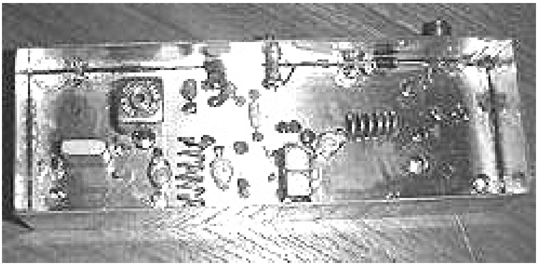
Unmodified Drake's measured at 2.5GHz RF and 200MHz IF showed that the units would comfortably meet their original specification if used within the intended band. Dave and I did many measurements of IF and RF response using sweepers, detectors and regular network analysers and concluded that the IF filter rolled off rapidly below 200MHz, and it also exhibited a marked dip around 145MHz! The RF filter also rolled off rapidly below 2.5GHz. These measurements led Dave to try some suggested modifications to the filtering which resulted in the noise figure of the test unit falling to around 3dB and the insertion gain climbing to a more comfortable 25dB. These modifications were published on G3PHO’s 'World above 1000MHz' web site:

www.g3pho.org.uk.

Essentially, these modifications involved lengthening the lines of the printed RF filter and adding one extra IF filter capacitor.

A further modification was to change the front end GaAs FET for an Agilent MGA 86576 GaAs FET MMIC. Although less straightforward than the filtering changes, the resulting improvement in noise figure is worth having. It should be noted, the length and quality of the grounding connection between the source leads of the MGA86576 and the ground plane on the reverse of the board must be short and clean (no excess solder) or the device will self oscillate - guaranteed! However, it was obvious that the noise figure could be further improved. At the 1999 Amsat-UK Colloquium Dave introduced the ceramic filter change. This involved replacing the existing, lossy, printed RF (image) filter with a small ceramic dielectric filter meant for 2.4 - 2.45GHz LAN products. When this filter was incorporated the noise figure of the converter fell to around 1.8dB, which is close to the claimed noise figure for the MGA86576 device. The converter also became more 'docile', possibly indicating less radiation loss from the printed filter, with it's always present potential
for instability.

With the modifications above the Drake 2880 is a very worthy converter for use with AO-40. However, there are still several pitfalls for the unwary. Dave Bowman, G0MRF has further



details of the Drake conversions on his web site:

www.g0mrf.freeserve.co.uk/

and may also have some of the conversion filters, MMICs and crystals for sale.

Beware the helix antenna

The MGA86576 is susceptible to input mismatch instability. One very popular UK produced multi-turn helical antenna has been measured with a very low input return loss of only a few dB. A Drake converter attached to one of these antennas will self oscillate. The use of several feet of lossy coax between the antenna and the converter will stop the oscillation, but is self defeating in so much as the overall system noise figure may then be unacceptable. The converter needs to be right at the feed point for optimum results.

The commonly accepted feed point impedance of the multi-turn helix of 140R is only obtained when the helix is centre fed1 . The more popular method of feeding on the periphery produces a much higher feed impedance requiring the use of a 1/4 turn matching section on the first 1/4 turn of the helix. This has to be very carefully tuned by bending the matching section towards or away from the reflector plate to obtain an acceptable match. A carefully matched section will start close spaced to the reflector and progressively further away at the other end.

A few further notes of caution must be mentioned when talking about helix antennas.

WA5VJB has made measurements on many of the popular G3RUH helix fed dishes in the USA. Very few of these have produced the expected results in terms of gain and circularity (and match?). Whilst G3RUH went to great trouble to get his design right for publication, it seems that many builders of these antennas don't follow the instructions. If you are using a helix fed dish and are not receiving the AO-40 S2 beacon very well when others are, it may be your dish feed! Also, the axial-mode helix is not always an axial mode! A long axial mode helix can behave like a normal mode helix (rubber duck) at some much lower frequency and may inadvertently receive high levels of lower frequency signals, especially since the axial mode helix is probably beaming into the sky and then looks and works for all the world like a vertical rubber duck antenna! A low loss, 2GHz high pass, filter between the helix and the converter may be a good idea in areas where there are lots of commercial VHF transmitters. This is especially true if you use a very wideband preamplifier in front of your 2.4GHz converter.

Converter supplies

It seems like the supply of Drake 2880 converters has, at least temporarily, dried up. I understand an alternative device has been found in the USA. However, in Eire, there is a similar 2.5GHz MMDS service to that in the USA. I have seen at least one design of MMDS converter in Eire that looks like it could also convert to 2.4GHz. I don't know if it has the same design PLL local oscillator to that in the Drake (many of these designs are similar) but maybe one of our Irish friends could investigate this possible source of converters?

Close

Receiving and analysing signals from the 2.4GHz satellites is both fascinating and easy, but it can be full of pitfalls for the unwary. If you don't get the results you expect, do check out what I have said above.

Programming a DMC 23GHz Brick Oscillator

Chris Bartram, GW4DGU

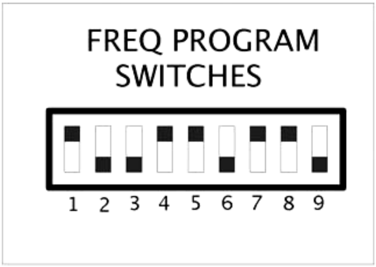
Despite my usual attitude to modifying surplus kit, I'm in the process of putting a 'quick converter' together using surplus bits in order to see if my 2.4m dish is accurate enough to use on 24GHz EME (and for some 22GHz radio astronomy).

A year or two ago I brought a DMC-110366 PLO brick, believing the seller that it might be useful as a 1.3GHz signal source. In fact the brick operates on around 23.5GHz and has good close-in noise performance and stability. After much playing with the DIP switches, I've managed to extract a very useful frequency from the unit: 23.616GHz, which is 24048-435MHz. As I've always used 70cm as an IF for 3cm, I'm rather pleased!!

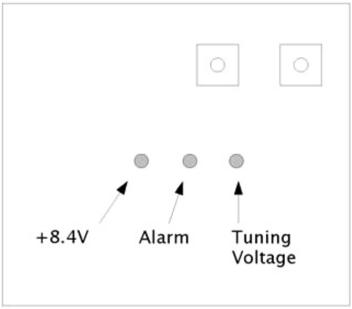
The brick requires a single +8.4VDC supply at about 1A max. There are alarm (essentially a lock detect) and tuning voltage outputs. The output level seems to be of the order of +8dBm at the 'XMTR LO' SMA output and around +4dBm from the 'receiver LO' output. I say 'seems' as I don't have a calibrated power meter and the spectrum analyser calibration is a bit suspect at 23GHz.

Alignment is a matter of setting the DIP switches and then adjusting the DRO centre frequency, using the only unmarked adjustment on the top panel of the unit.

Monitoring the tuning voltage output should show the effects of tuning the DRO. The best setting is when the tuning voltage is at the centre of its limits.



Location of DIP switches on DMC-110366 23GHz LO Module for 23.616GHz Output (435MHz IF = 24048MHz)



DMC 23GHz LO Power and Control Connections

Some DMC ‘23 Whitebox’ information

Steve, G8JMJ

Editor’s note: The following plea for information appeared on the UKuG Internet Reflector:
“Does anyone have any information on the various modules within the DMC "white boxes" for 21/23GHz ...
are they any use ?“

The previous pages describe GW4DGU’s findings regarding the oscillator contained in the equipment but here’s some more information, this time from Steve, G8JMJ. who posted this on the reflector in reply to the query above.

The DMC-110366 oscillator brick uses some form of a dual loop, but I’ve not had the time to analyse it properly yet.
A few points:

- The unit has two crystals, a 14.000MHz (TO5 can style) PCB mounted and what appears to be another one buried on the PCB below, contained in a small oven (will confirm and measure it's frequency)
- The sapphire trimmer on the top of the unit trims the 14.000MHz oscillator and can be used for fine frequency adjustment. The other oscillator can be adjusted by removing the large screw-in plug on the side of the module.

The bit switch positions have the following step sizes.

1 =3D 1.75MHz
2 =3D 3.50MHz
3 =3D 7.00MHz
4 =3D 14.00MHz
5 =3D 17.5MHz
6 =3D 35.00MHz
7 =3D 70.00MHz
8 =3D 140.00MHz
9 =3D 280.00MHz

The big toggle switch buried inside the module (accessible through a hole on the top) switches the DRO into free-run mode to allow for frequency setting with the loop open.
It should be noted that, although the 'synthesiser' circuitry can step these relatively large steps, the DRO cannot and so will require manual retuning before 'locking' occurs. Not all codes are valid. I have had the unit locked where the LSB (1.75MHz) step switch is in-operative.

On the other bits, the LNA runs from +8.4VDC as does most of the entire box. The gain at 24GHz was only about 17db. I’ve not measured NF yet.

The TX up converter works reasonably well at 24GHz with an IF of 200 to over 1300MHz (obviously lower conversion gain at the top end).

The TX image filter will retune so that the top end of it's response is at about 24.1GHz, insertion loss about 1.5db but with about 1db of ripple. Bandwidth, from memory, is about 300MHz.

Useful 10GHz Filter with 10MHz Bandwidth

David Wrigley, G6GXK

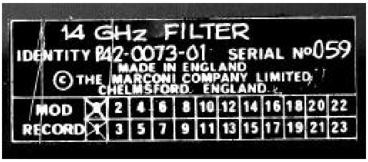
In the past have had some problems with my 10GHz transmitter. In order to ensure that I was peaking up the right signal I thought that it might be useful to feed the output through a narrow filter before measuring the output power. There are, of course, many such filter designs (ref. 1) in the Microwave Handbook and there is also the G3KNZ program (ref. 2) which permits you to design your own filter, as the following example shows:

Waveguide Band-pass Filter Design using WG16
Filter type B,
Ripple= 0 dB, Frequency= 10368.1 MHz, BW= 10 MHz
WG16 Internal Dimensions = 22.86 x 10.16 mm
Material=CU, Skin depth= 2.237868E-02 mm, Iris Thickness= 1 mm

Waveguide Filter Dimensions:
Cavity No: 1 = 18.2 mm (less 5% = 17.3mm actual to allow for tuning screws)
Iris Hole size: 0 = 6.4 mm
Iris Hole size: 1 = 6.4 mm
Average Unloaded Q of Cavities = 5500
Insertion Loss = 1.64 dB
Lg/8 Matching screw spacing= 4.7 mm

I was searching for some scrap copper waveguide in order to make just such a waveguide filter, when my eyes fell on a discarded **14GHz** filter which was originally part of a Marconi 14GHz up-converter, many of which were sold off at meetings last year. The filter had been discarded because it was originally thought that it couldn't be re-tuned to any useful frequency. You know the feeling - it looks so nicely made that one feels instinctively that it will come in useful one day!

A further examination revealed that it was indeed a very useful piece of microwave kit. The filter is a six stage unit built up of separate silver plated cavities interleaved with copper iris plates and the whole thing bolted together by means of four longitudinal bolts. The cavities are basically very rugged sections of WG18 and the cut - off frequency for that is about 9.5GHz. So, it should be OK at 10GHz. The thing to do is to remove some of the iris plates and calculate what the new cavity length is for 10368MHz, but first we can, out of interest or curiosity, check that the original design is what we thought it should be and get some experience using the Steve Davies, G3KNZ filter program running under GWBasic.



14 GHz FILTER
IDENTITY P42-0073-01 SERIAL No059
MADE IN ENGLAND
© THE MARCONI COMPANY LIMITED
CHELMSFORD. ENGLAND.

MOD	☒	2	4	6	8	10	12	14	16	18	20	22
RECORD	☒	3	5	7	9	11	13	15	17	19	21	23

Fig 1: This is the self adhesive label from the original filter unit

Using components of old 14GHz unit Original design:
WG18 section 15.8 by 7.9mm
Cavity length 12.35mm
Iris Dia 5.7 on outer units, all the rest 5.1mm
Iris thickness 0.3mm
Design frequency marked on front panel, 14275MHz
Several runs were made, adjusting the bandwidth until the iris diameters came out right.

Running the program:
Filter type B,
Ripple= 0 dB, Frequency= 14275 MHz, BW= 50 MHz
WG18 Internal Dimensions = 15.8 x 7.900001 mm
Material=AG,Skin depth= 2.554082E-02 mm, Iris Thickness= .3 mm

Waveguide Filter Dimensions:
Cavity No: 1 = 13.4 mm (Actual cavity = 12.35 , therefore using a tuning allowance of 8% on calculated)
Iris Hole size: 0 = 5 mm (Actual holes mainly 5.1 mm in 6 cavity filter)
Iris Hole size: 1 = 5 mm
Average Unloaded Q of Cavities = 5033
Insertion Loss = .49 dB
Lg/8 Matching screw spacing= 3.5 mm

Revised cavities for 10368.1MHz Design Filter type B,
Ripple= 0 dB, Frequency= 10368.1 MHz, BW= 10 MHz
WG18 Internal Dimensions = 15.8 x 7.900001 mm (cut-off freq 9.5GHz)
Material=AG, Skin depth= 2.176689E-02 mm, Iris Thickness= .3 mm
Waveguide Filter Dimensions:
Cavity No: 1 = 34.3 mm (less 8% allowance for tuning = 31.5mm)
Iris Hole size: 0 = 6.6 mm (Used 5.7mm)
Iris Hole size: 1 = 6.6 mm
Average Unloaded Q of Cavities = 4646
Insertion Loss = 1.94 dB
Lg/8 Matching screw spacing= 9 mm

Conclusion

For a 10368MHz filter, the desired cavity is 31.5mm long. Divide this by the current length of a cavity block (12.35) and we get 2.55 blocks. So just over two and a half blocks are required for the new cavity. One was cut in half fairly easily using a junior hacksaw with a sharp new blade and the cavity block held in a vice. The sawing marks were removed from each piece using a broad smooth file. The cavity blocks were then rearranged to form the 10368GHz filter. Tuning up a single cavity is fairly straightforward provided one has a signal source and a detector. I used a Gunn Diode oscillator and diode detector. The Gunn Diode was first set up to 10368MHz using a frequency counter, then the filter was peaked up using a diode detector. I later tried to check the pass band of the filter unit by means of manually sweeping the Gunn Diode oscillator with a frequency counter on the output, but this was difficult due to the very sharp tuning which caused the frequency counter to lose lock either side. What was clear was that the filter had a bandwidth of a few MHz either side of 10368MHz. I found that both tuning screws in the cavity had an effect and had to be adjusted for maximum output.

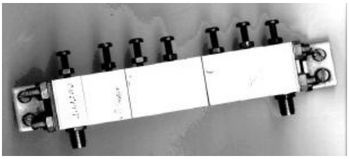


Fig 2: The finished filter

you can figure out the posit ions of the half blocks by means of the gaps in the tuning screws
Clearly many filters from 10GHz to 18GHz could be constructed from these bits – a sort of universal
filter kit and it will no doubt be useful in the future .... not bad for an initially rejected item!

References

Ref 1 Examples of practical Waveguide filters 1995 reprint Microwave Handbook Vol. 3 page
18.18. (G3JVL)
Ref 2 Waveguide filter design program 1993 reprint Microwave Handbook Vol 2 page 12.26
(G4KNZ)

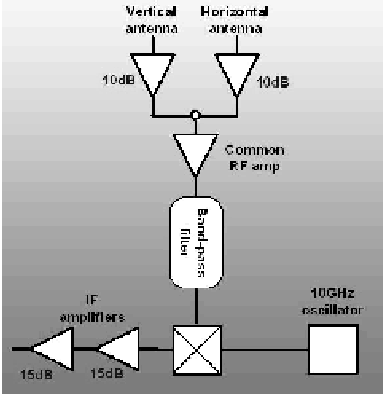
© David Wrigley G6GXK

A (re) introduction to 10GHz via modified Satellite LNBs

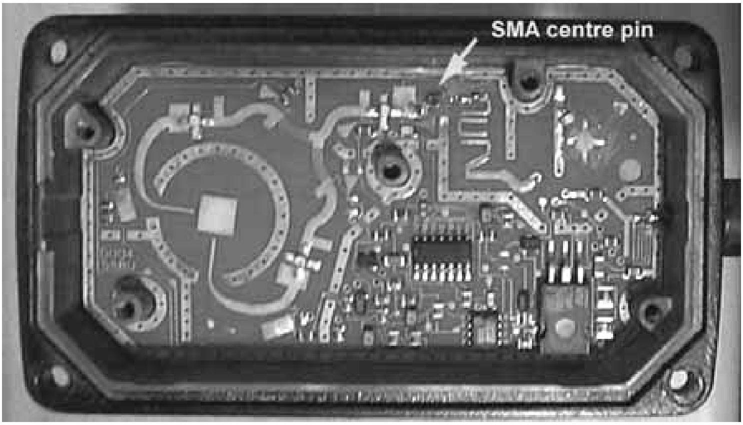
Bernie Wright, G4HJW

Dropping in late to the Lincoln Hamfest a few years ago, I found one of the traders selling off a box of about 30 satellite LNBs for £10.00, and just couldn’t say no. Most LNBs are as shown here to the right - the older ones may have more stages and an active, rather than diode, mixer but the block diagram shown should still be quite representative of what is currently around.

The MMIC IF stages were an obvious candidate for re-use, and it was straightforward enough to convert the stages to provide amplifiers with a gain of 40dB or so from HF to a couple of GHz. With these, built onto the back of a dipole/reflector plate combination, noise free signals from the local 23cm TV repeater could be piped down through long lengths of pretty “ropey” coax without having to worry about signal loss.



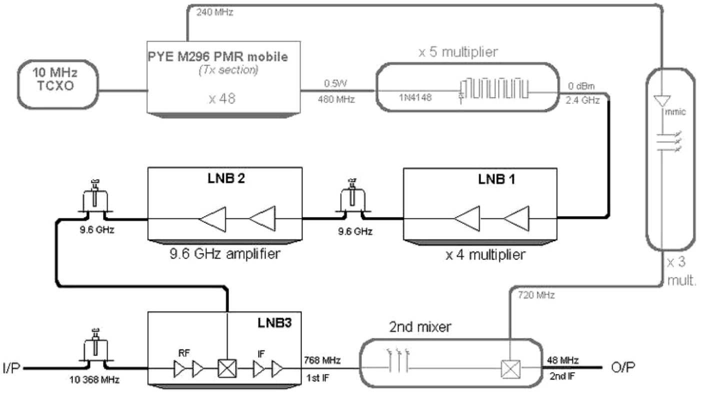
It was a year or so later that another use came to mind. Interest in SETI had rekindled a general interest in radio astronomy for several of us and the notion of producing an 11GHz interferometer, using two 60cm dishes with their LNBs, began to appeal. We knew that the RF stages of most LNBs of the time gave about 10dB gain, and that there were always at least 2 stages, so early on we thought about getting in after the second stage on each dish mounted LNB, bringing the feeds out to a combiner (OK, just a



BNC T piece!) and feeding the result in to a third LNB, which would only be modified by fitting an sma socket in place of integral antenna probe. The loss in even "bog standard" quarter inch coax over a few metres would not approach the 20dB gain figure of the RF stages. It was soon clear that, with many of the LNB types, you could simply drill through both the pcb and the aluminium housing it was fitted to, after the second RF amplifier, You would then be able to open up the casting hole and sweat an sma socket onto the ground side of the pcb. It turned out to be very easy to do. On LNB types where part of the casting did get in the way, a hacksaw and file soon resolved the problem.

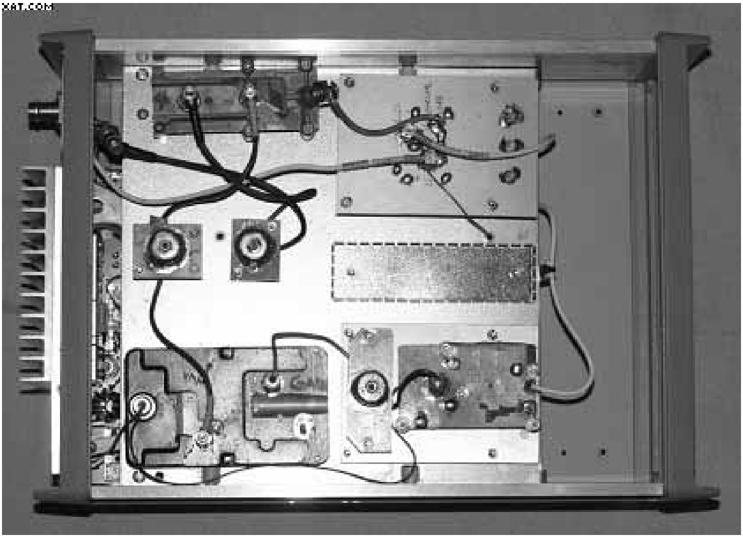
The 1-2 GHz IF from the third LNB was fed via a couple of further MMIC stages into a diode detector. After much adjustment (and fun), good interference patterns were obtained from sun transits with this arrangement. During the following year, the idea of replacing the ceramic resonator based LO with something more stable grew. It had been 20 years since my last experience with 10GHz (and those free running 723A/B klystrons) but I at least knew that narrowband operation had now made klystrons and Gunn diodes (and Barrett diodes, if anyone remembers those) all but obsolete.

About half of the surplus LNBs had a discrete diode pair mixer, with MMIC IF gain stages and a single GaAs FET oscillator. Again, it was easy enough with most of these to drill a hole through both the pcb and casting at the point where the FET oscillator fed the diode mixer and fit another sma connector so that an external crystal multiplier LO could be applied. It took a while to figure out a suitable multiplier, but, in the end, a Philips FR5000 PMR 1W VHF driver module was used to drive a pair of 1N4148 diodes. A 12 stage(!) pcb interdigital filter followed it to give a very clean x6 multiplication up to just over a GHz. Some 'Heath Robinson' MMIC amplifiers and pipe cap filters then took the LO up to 10GHz. I can't remember now what the IF was but it was single conversion to something around 150MHz. With this converter, fed from a second LNB just used as an antenna-with-integral-preamp (one of the interferometer units), a trip down the M11/A130 soon had GB3CMS coming in at good strength on the road side at Ford End. The real downfall with this converter was the poor stability of the crystal oscillator – but it was a start



and the bug had bitten. The frequency stability problem eased when it was realised that 10MHz, 12.8MHz and 14.4MHz TCXO’s – all readily available, would multiply up to provide useful LO frequencies. Since 14.4 MHz also multiplies up to exactly 10 368MHz, this later became the basis for a CW/FM transmitter [likewise a 12.8MHz TCXO was/is being used as the basis for a low power beacon in the Cambridge area].

The current 10GHz converter here uses a 0.5ppm 10MHz TCXO in a double conversion configuration.



The first IF is 768MHz [9.6GHz LO] and this is brought down to 48MHz using a second LO of 720MHz, all driven from the 10MHz oscillator as below: A more practical approach to the LO multiplier at 480MHz, for anyone who has access to an ex Ionica head, would be to use the duplexer to select the 3.2GHz harmonic . These items have a surprisingly wide tuning range, and both RX and TX sides will cover 3.2GHz (9.6GHz LO) or 3.456GHz (10.368GHz). They can also be operated end-to-end from the TX to the RX port, giving more/sharper selectivity.

Liberal use of surplus satellite LNBs was [deliberately] made in this converter, though no attempt was made to use the integral image filter as part of the 9.6GHz LO multiplier. However, this did work well on the transmitter strip to multiply 3456MHz up to 10.368MHz, giving 40dB plus rejection of all other harmonics (without any modification to the etched filter). Instead, a couple of 15mm pipe cap filters were used.

These, together with the single image pipe cap filter, can be seen in this photograph of the converter: Yet to be tried is modifying the LNB oscillator so that the stage acts as both buffer and selectivity block using the integral ceramic puck. Has anyone already tried this approach?

The single image filter seems to have been quite adequate with such a high 1st IF, and the through loss not an issue when fed via a remote LNB antenna-with-integral-pre-amp arrangement.

Incidentally, RG223 seems to have a

loss of 4 to 5 dB/m at 10GHz, so up to 2 or 3 metres of interconnection lead between antenna and converter is OK. For longer lengths, such as chimney to shack use, it just needs another LNB configured as a 20dB amplifier added as a line amp.

Two LNBs are used in the LO path (laziness rather than necessity) configured as 20dB amplifiers. The first takes the 2.4GHz feed at about 0dBm, ensuring that this amplifier clips. The fourth harmonic is selected by the pipe cap filter and fed to the second LNB for amplification. About 10dBm is then available, so another pipe cap filter was put in for good measure prior to feeding the third LNB, taken from a mini-dish and fully used as the 10 368 to 768MHz converter stage. The early Cambridge mini-dish LNBs seem to work very well (see photograph on the next page) and the converter is fed from another one of these.

Laziness dictated the use of a separate mini-dish for transmit. With 300mW of CW, there has been no need to power down the receive LNB when on transmit. Since there is already a very noticeable increase in background noise as the dish is dropped down to the horizon, no attempt has been made to optimise the LNB probe feed at 10 368MHz.

GB3CCX at 150km is pretty much always copyable with this set up and thermal noise from nearby trees and buildings can be noticed as the antenna is rotated. So all in all, the LNBs have proved to be a good buy.

Every type of unit that was modified to act as antenna/ pre-amp units had a noise figure that were good enough to detect ground noise easily. Some of the mid 90’s (analogue) Cambridge units had three cascaded stages following the antenna, but most have only two. A good rule of thumb is to expect 10dB of gain per stage. Using a hot air gun to remove devices has proved reliable.

Careless handling after that point has destroyed a few devices through static damage, but that is all The only comment I would make is that current limiting on the GaAs FET stages never seems to be provided, so that shorting the gate to ground always causes device destruction.

As for other LNB uses – I wonder if anyone has attempted using dual output units, which have the RF section duplicated, as the basis for a phasing transmitter/ transceiver?

HS400 - A useful and simple modification

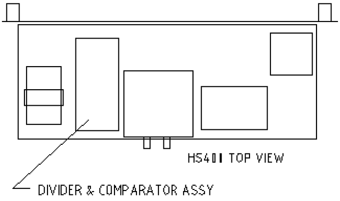
David Wrigley G6GXK

The HS400 contains a very useful and high quality reference oscillator. We all need a frequency source that we can rely on to set up the local oscillator in our transverters and any frequency counters that we may have. This 5MHz reference oscillator is just the job to use as a reliable local source. It was the author's intention to control this oscillator by synchronising it with MSF – hopefully more of that later. The first stage is to get at the voltage control connections and the output. These are all available on the connector at the end of the Divider and Comparator PCB which is located next to the reference oscillator.

A simple connection would be possible but unwise. These oscillators can be affected by load even when buffered by a single stage and it is vitally important to keep it low and constant.

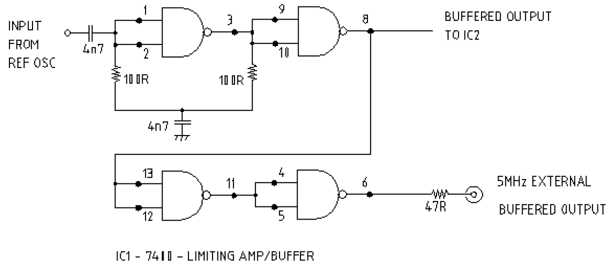
It was decided that the connection would be made further away from the oscillator and the output of the Limiting Amplifier would seem appropriate. With amazing good fortune there are two spare gates in the 7400 (IC1), which have been used in this modification to further isolate the output from the internal circuitry. The only further consideration is the possibility of shorting the output. Well the man who doesn't make mistakes is a myth – we all do. So I have added a 47R resistor in series with the IC – this also cuts down the output voltage swing to a more reasonable level and limits the potential for damage to other equipment. Well that's it. The diagram gives a bit more detail.

The next stage is to build a PCB to



1. LOCATE THE DIVIDER & COMPARATOR PCB
2. DETACH THE CONNECTORS & REMOVE THE PCB
3. LOCATE THE 7400 IC AT THE END OF THE PCB
4. LINK PINS 8,12,13 TOGETHER
5. LINK PINS 4,5 & 11 TOGETHER
6. CONNECT SCREENED WIRE TO PIN 6
7. CONNECT SCREENED WIRE VIA 47R TO NEW BNC CONNECTOR ON REAR PANEL
8. REFIT THE PCB AND CONNECTORS

(c) David Wrigley April 2010



IC1 - 7400 - LIMITING AMP/BUFFER

(c) David Wrigley April 2010

control this oscillator from MSF via a PIC chip controller feeding a Digital to Analogue Converter IC.
The purpose will be twofold.
1. The provide a long term lock to MSF accuracy
2. To minimise drift in the short term between MSF updates.
It should be possible to get an accuracy of 1 part in 10^10 using this system which is 1Hz in 10GHz. This might be bit over the top in terms of

HS400 – Electrical Frequency Control of the oscillator

David Wrigley G6GXK

The HS400 contains a very useful and high quality reference oscillator. We all need a frequency source that we can rely on to set up the local oscillator in our transverters and any frequency counters that we may have. This 5MHz reference oscillator is just the job to use as a reliable local source. It was the author's intention to control this oscillator by locking it to MSF or another reliable source and these modifications are intended to allow us to do that. The previous modification showed how easy it was to get out a buffered 5MHz output to a socket on the back panel. This modification allows us to electrically fine tune the crystal frequency and will ultimately allow us to set it to 5MHz manually (and later automatically) to within a few parts per 10^10.

Anyone who has seen the schematic diagram for the HS400 will have noticed that a provision for electrical frequency control is in place. How strange then that it doesn't work! The reason for this is that whist the external wiring is in place; Toyocom does not fit the internal components into the reference oscillator.

To fix this problem we will need to get inside the reference oscillator can and make connections to the internal stabilized supply and then fit a Varicap diode and series resistor to the crystal circuit.

One Varicap? – well, I believe that using only one diode can theoretically produce some phase noise and that two diodes back-to-back minimise this effect. However, this 5MHz reference oscillator using only one diode has been used as a source for the Manchester 10GHz beacon and no significant phase noise has been heard in the receiver. I am reluctant to put in extra components unless I've experienced the need for it.

Schematic correction

Before we get deeply involved, it is a good idea to check the cable between the reference oscillator and the adjacent PCB. In the two units modified so far this cable did not agree with the circuit diagram. The circuit diagram should be:-

The Orange wire from SKT C/pin2 connected to SktA/pin 4 (not 2 as shown)

The Blue wire from SKT C/pin7 connected to SktA/pin 2 (not 4 as shown)

Getting to the oscillator PCB

You will need to remove the oscillator can from the HS400. The procedure is

1. Turn the top panel screws a quarter turn (opposite ways!) to release the top panel.
2. Locate the Reference oscillator can at the opposite end to the mains transformer.
3. Unplug the connector and slacken off the two clamp fixing screws to release the can.
4. The can will then slide out towards the front of the HS400 unit. Move the can over to a clean area of bench and have a small tray handy to hold the collection of screws to be removed.

To strip the unit,

1. Unfasten the four small cross-point screws (with washers) on two sides of the can.
2. The base and assembly will then be able to be pulled out by means of the projecting screwed rods.

3. Unwrap the insulating fibre material and put this to one side.

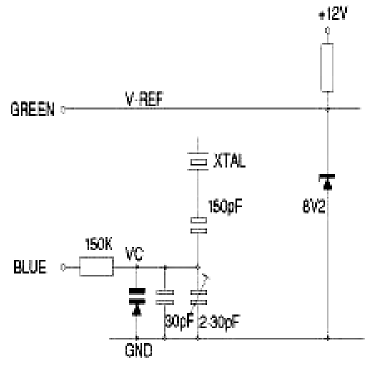
4. The PCB in which we are interested is the small top one and can be worked on without any further dismantling at this stage.

The modification,

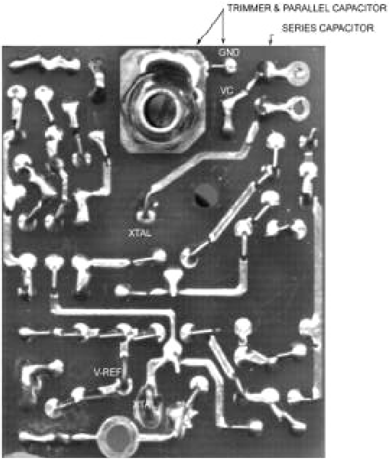
1. Fixing the Varicap diode and associated capacitor and resistor.

The Varicap diode used by the author was a small plastic (SOD323) 17pF (at 0V) type. The types likely to be suitable are the Farnell BB149A (316-2837) or alternatively (in SOT23) the Farnell, BBY31 (300-0310) or BBY39 (316-3076).

The Varicap diode is soldered across the points marked “GND” and “VC”. Also from “VC” is connected the (short) end of a wire ended 150K resistor. The other end of this resistor is connected to a wire and insulated with sleeving. The other end of this wire is soldered to the blue wire termination pad on the lower PCB. This wire provides an external connection for the frequency control voltage.



TOYOCOM 613 REF OSC - PARTIAL SCHEMATIC DIAGRAM



Useful 1GHz Oscillator Module

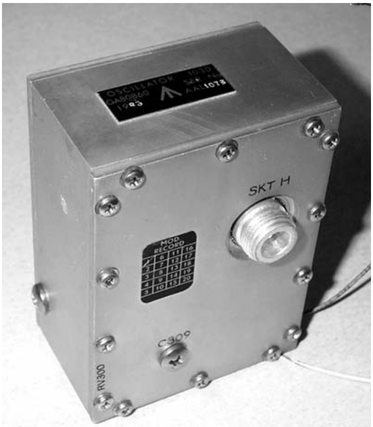
Gordon Fiander, G0EWN

Some of you may have noticed some extremely well constructed units for sale at various recent UK rallies. They are normally being touted by dealers as 'ovened oscillator units' with output at 1030MHz and are selling for around £2.50. I bought a couple at an Autumn rally at Wakefield.

On checking the units at home, the following became apparent. First the units are extremely well built; the milled brass enclosure alone being well worth the cost! (See photo). However, the units are not 'ovened'. Inside, a crystal oscillator (on 85.83333MHz) is followed by an amplifier stage, which in turn drives a diode multiplier feeding a tuned output line. The output line is tuned to pick off the 12th harmonic, in this case giving about 18dBm out on 1030MHz.

I used the bench supply set at +13.8V to test the units--both drew about 40mA. Having ascertained the units worked OK I fitted a 5th overtone crystal on 86.4MHz to one unit. The unit worked fine without need to retune, giving an output on 1036.8MHz; the 10th harmonic of this falling in the 10GHz band. The note was quite stable after a short period from turn on , even when listening on 10GHz, and the crystal could be easily adjusted to give a signal at the band edge or around the calling frequency on 10GHz. I intend to fit a DB6NT heater for even better stability and hope to be able to use as a marker to check RX (frequency?) when out portable.

The units need a stable supply as small voltage changes affect output frequency. I hope to change the second unit I have by fitting a crystal on



90.666 to give output at 1088MHz for a LO for another project; it might also work with 96MHz to provide an 1152MHz source and as such are a good buy.

Note that you will need a hot iron to desolder the existing crystal as silver solder appears to have been used throughout. Also if the supply drops below +11V the oscillator stops (it may have originally had +20v supply?). I have made a simple diode marker generator to go with the unit and can hear the signal throughout the shack. When the dish is pointed at the unit the signal is 59++

Surplus

Test Equipment

An External Mixer for the HP8555A Spectrum Analyser Plug in

Chris Towns, G8BKE

This article describes an extension of this useful instrument to 24GHz and above, by the addition of a simple add on mixer.

In its standard form, the 8555A plug-in for the HP141T Spectrum Analyser will accept inputs in the range 10MHz to 18GHz via the input N type connector. There is a modification available to allow access to at least 24GHz via this connector but 24GHz via an N type is not really recommended. Indeed, I believe the modification precludes one from using the external mixer capability.

The 8555A will allow the use of an external HP mixer via a front panel BNC socket and this will extend the range of the instrument to 43GHz, just short of the amateur band! To obtain displays at this sort of frequency the 8555A LO is automatically multiplied in the external mixer and mixed with the test signal. In conversation with Brian, GM8BJF it turned out that he was using a "home-

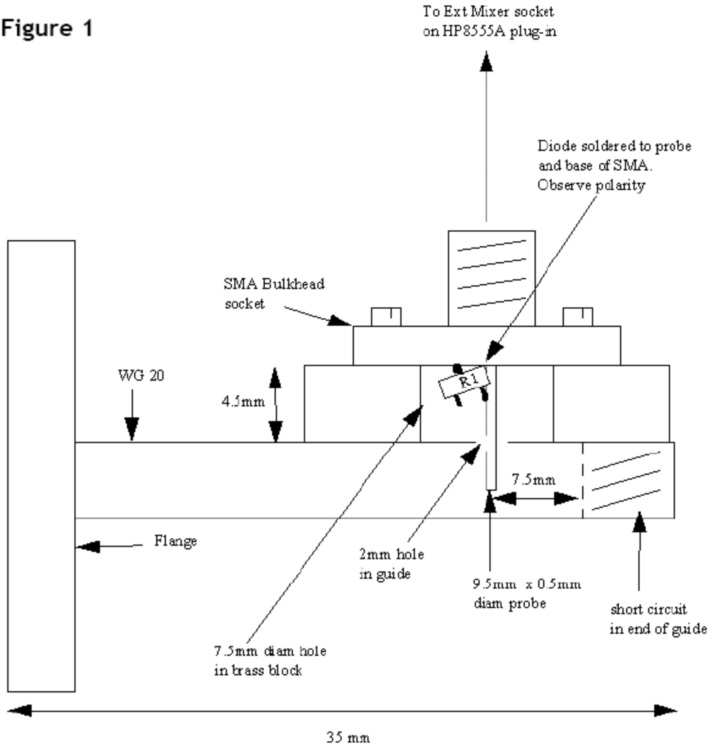


Figure 1

brew" external mixer to look at a 24GHz signal quite successfully and its was decided to try this approach to look at 47GHz. Although the amplitude calibration using this method is not really absolute, adjustments for maximum output, sideband selection and signal cleanliness can nevertheless be examined.

With no genuine surplus HP external mixers available, a small unit would have to be made as GM8BJF had done. However, without access to a 1N26E diode that Brian had used, experimentation with more readily available diodes was needed.

Since the mixer was going to be used at 24GHz and above, it was decided to make the unit in a very short length of WG20 which was to hand.

Figure 1 shows the outcome of using this approach.

The SMA socket was mounted on a small 4.5mm thick brass block soldered to the guide broad face, with a ~7.5mm diameter. hole in it to accommodate the mounting of the diode. A short circuiting brass block was soldered in to the end of the guide at the position shown. The probe from the SMA protrudes into the guide via a 2mm hole in the guide wall. The probe itself can be a wire extension of the SMA pin, or better still the SMA pin itself.

A number of diodes were tried including those out of LNBs. All worked well, but, in order to obtain a readily available source, the commonly available HP HSMS 8101 diodes which are available from Farnell (Stock No. 994-649) were used. Only really applicable for use in a surface mount applications their packaging is not ideal, but since maximum efficiency is not required, they seem quite satisfactory here. Care should be exercised in soldering in the diode as static can damage the device. Correct orientation of the diode should also be observed, since a positive bias voltage is applied to it from the 8555A. There are three leads on the diode package, but only two are used so it should be mounted with the lettering as shown, ensuring that it is correctly orientated. It is quite possible that other diodes such as the DDC4561 or even a 1N21, could also be used, but these have not been tried. The external mixer arrangement on the HP 8555A is clever in that the DC bias, IF and LO all share the same cable to the mixer. Thus if a short length of good SHF cable is used, suitable for use at 2-4GHz (e.g. SUCCOFLEX) no difficulty should be experienced in obtaining results. Adjustment of the "Ext. mixer bias" pot on the 8555A optimises the mixer /multiplication process for best signal into the analyser.

Since the frequency dial on the 8555A runs out at 43GHz, some other means has to be employed to determine the frequency of the wanted signal. Luckily HP also put the LO frequency on the top of the scales. Thus one knows what LO frequency the mixer is seeing! The IF centre frequency on the highest ranges of the 8555A is 2.05GHz. Knowing this, one can, by some arithmetic, work out what the LO frequency should be. As an example, to display a 47.088GHz signal, 47.088 minus the 2.05GHz IF will require an LO of 45.038GHz. This is obtained by a x12 multiplication of 3.75GHz LO in the mixer. Thus the frequency set pointer should be set as close as possible to the LO frequency of 3.75GHz on the top scale of the instrument.

Although I have not tried this, due to lack of a good 3GHz counter, it should be possible to connect a counter to the ext. mixer socket (watch the DC bias!) or the first LO output socket and with the 8555A set to "manual sweep" set the LO to precisely this frequency and then reset to "Int. sweep mode" again with the span set to say 1MHz. However, having said this, it appears that the "Signal Identifier" on the 8555A still operates in this non-standard mode thus one can check by the normal means if the signal being displayed is the correct one bythe "usual two divisions to the left" offset.

Note that for the external mixer to function and for the LO to be directed out of the instrument correctly the 8555A must be operated with the band selector set to one of the frequency ranges above 18GHz.

Also note that in this mode the "Input Attenuator" of the 8555A is not functional, although the "Log Ref. Level" control is. If the power of the signal has been measured previously, on a power meter, then a crude calibration of the vertical scale can be made. The signal being checked into the mixer should be kept to 2mW or less as driving it harder only produces more mixer products and tends to confuse the measurements. Keeping the input to this sort of level also ensures the longevity of the diode!
It is hoped than when I'm operational on 76GHz that the same method can be used to examine the signal at that frequency.

A similar diode mounting arrangement is being examined with a view to producing a useful waveguide noise source for noise figure measurements above 24GHz.

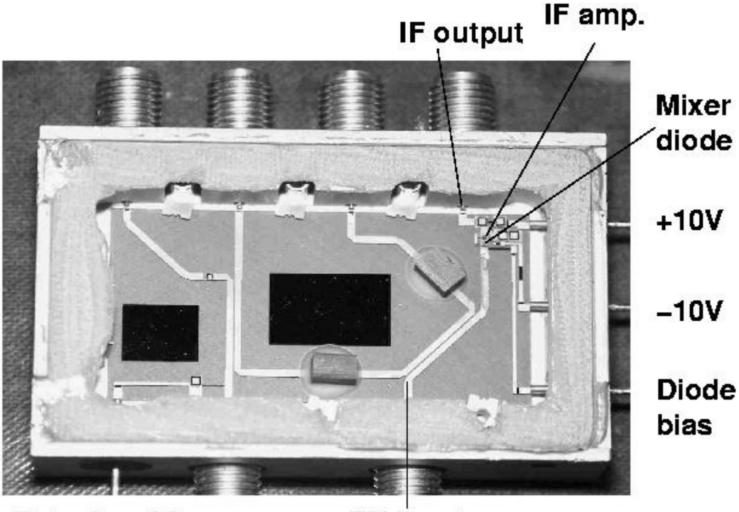
Repairing the Input Mixer of the HP8555A Spectrum Analyser Plug-in

Brian W, Flynn, GM8BJF

Recently I noticed that the sensitivity of my much prized HP 141T/8555A spectrum analyser had dropped rather drastically. I feared that the input diode had died for some reason and a d.c. check of the mixer diode, as described in the service manual [1], confirmed my worst fears. The diode was short circuit.

After a period of dejection I considered methods of repair. The first mixer is a thin film microcircuit which HP replace as a unit of it is damaged. There is also a small plug-in PCB assembly which contains select-on-test resistors, used to set the mixer bias and the 50MHz IF amplifier gain, to define the calibration of the unit. This PCB is changed along with the mixer. Deciding that there was little to lose, I removed the lid from the mixer package. The mixer substrate is housed in a gold plated metal case with a lid which is held in place with conducting epoxy adhesive. The lid can be removed by carefully prizing it off using a small screwdriver as a chisel while holding the mixer unit in a vice. This revealed the circuitry on the substrate shown in **Figure 1** below. The mixer diode is a chip device which is bonded to the substrate and the post mixer amplifier is a grounded base bipolar silicon transistor. The mixer diode output drives directly into the emitter of this transistor, presumably to give a good broad-band match to the mixer diode output impedance.

Initial attempts to replace only the mixer diode with a surface mount



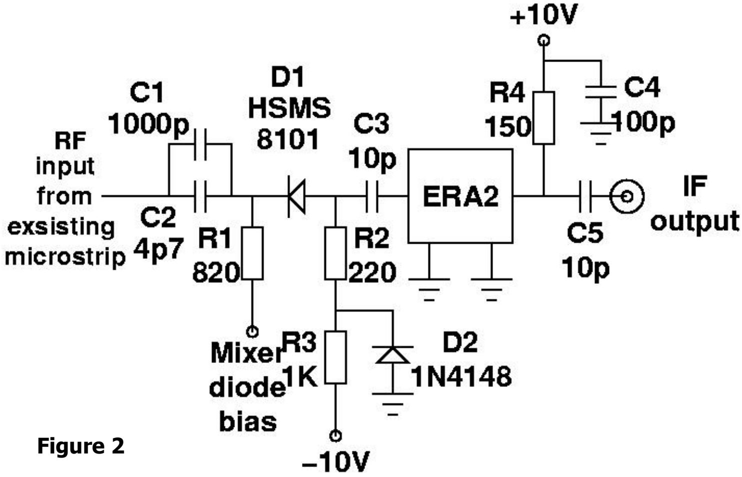


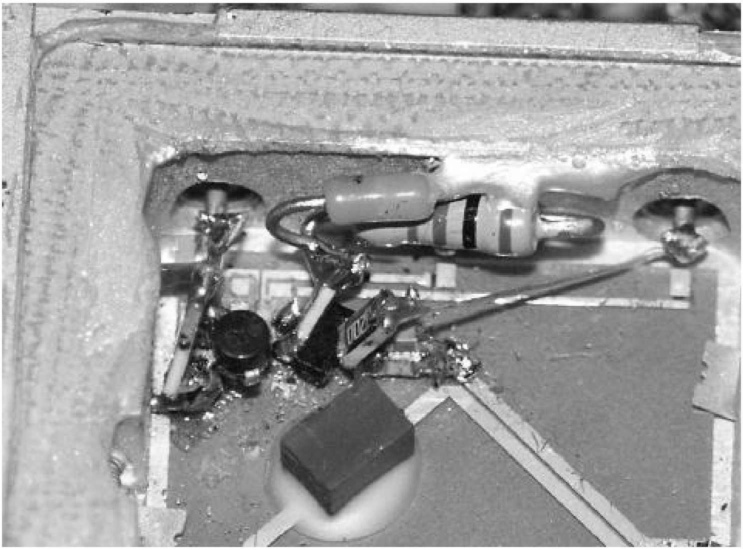
Figure 2

packaged device were unsuccessful. For this to succeed wire bonding facilities would be necessary. The next attempt was to replace all of the active circuitry as per the simplified circuit diagram shown in Figure 8-12 of the service manual. This was not very successful as the sensitivity dropped drastically.

The alternative approach adopted was to replace all of the active circuitry with the circuit shown in **Figure 2**, fabricated with surface mount components. The first step was to ground the two common leads to the ERA2 to the earth point on the substrate. This was achieved by careful soldering with a fine bit. Once the ERA2 was in place, the bond wires to the package connections were broken so that connection could be made to them. The ERA2 amplifier operates as a IF 2GHz pre-amplifier and provides a 50 ohm termination for the mixer diode. The HSMS-8101 is a Schottky barrier mixer diode which is available from Farnell. The capacitors C1 and C2 could be omitted if use can be made of the existing component mounted on the input microstrip. Attempts to solder to it were unsuccessful and resulted in its destruction, hence the use of C1 and C2. These were paralleled to give a broad band capacitor as the instrument operates from 10MHz to 18GHz. It would be preferable to retain the original part in circuit to maintain the response of the instrument at the lower frequency end of its range. The resulting modified mixer is shown in Figure 3. The result is not pretty, but it meant that the unit became usable again, (even if HP/Agilent would not approve of the accuracy!). The ground connections to C4 and D2 were made directly to the metal case with conducting epoxy. They were first soldered in place, in contact with the case, using the solder connections to locate them. The ground connections were then formed with sparing "blobs" of conducting epoxy.

Mixer Bias

The mixer is operated as a harmonic mixer to get the wide tuning range. It operates up to the fourth harmonic of the 2 – 4GHz YIG LO. To optimise the conversion loss the dc bias applied to the diode is set for each harmonic. This is controlled by fixed resistors



mounted on the daughter board assembly A10, which is mounted on plug-in board A6. Experimentation with the resistor values can improve the sensitivity. This was achieved by temporarily replacing the appropriate resistor with a 4k7 variable and varying it for best sensitivity on a suitable signal and then re-fitting a suitable fixed resistor.

Performance

The repaired mixer gave reasonable performance over the full range of the instrument. The amplitude calibration is not as good as it once was, but is within 2-3 dB on all ranges. To complete the calibration the resistors controlling the IF amp gain on the A16 assembly would repay attention. When time permits these will receive attention. After testing, the lid was replaced using more conducting epoxy.

Conclusions

This procedure was able to bring an otherwise useless piece of equipment back to life and allow it to be useful for amateur work. Manuals for the 141T mainframe and most of the plug-ins can be downloaded from the website:
http://www.logsa.army.mil/etms/find_etm.cfm
and can also be found at:
http://recordist.com/pdf
Unfortunately, the full manual for the 8555A is not publicly available on these sites but a nonetheless useful manual giving drawings and component locations and values can be had. The full operation and service manuals for the other 141T plug-ins, preselectors and tracking generators are available. There are also manuals for other items of test equipment available if you search the site.

Be warned they are large files!

References

1. Spectrum Analyser RF Section 8555A, Operating and Service Manual. Hewlett-Packard Company, Santa Rosa, USA, 1971.

Acknowledgments

The author wishes to thank Peter, GM4DTH for taking the photographs.

Parts listing

R1 820 0805
R2 1k 1/8W axial
R3 220 0805
R4 150 1W surface mount Farnell 507-878
C1 1000p chip
C2 4p7 ATC100A
C3 10p chip
C4 100p chip
C5 10p chip
D1 HSMS8101 Farnell 994-649
D2 1N4148
IC1 ERA2 Mini-Circuits

A New Pointer Drive for the HP8555A Plug-in

Chris Towns, G8BKE

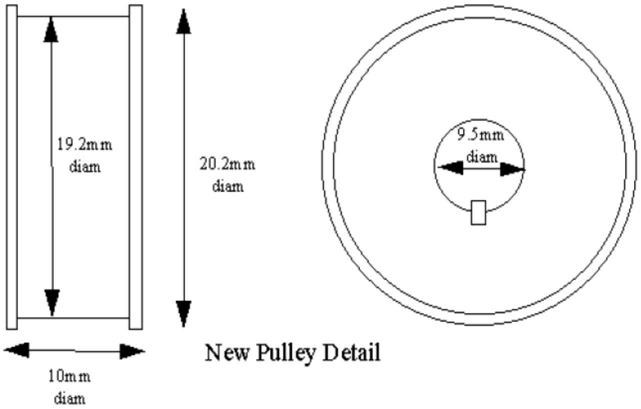
There are a large number of Hewlett Packard HP8555A 0.01 to 18GHz plug-ins for the HP 141T spectrum analysers around in both amateur circles and on the surplus market. One drawback of these units seems to be that the pointer drive mechanism, comprising a thin fibre belt, can wear with age and eventually snap. This renders the instrument pretty useless and to replace the belt is quite costly.

The following pages offer a cheaper solution which seems to be just as satisfactory as the original drive.
It involves replacing the cogged belt drive with a chord and pulley drive. Access to a small lathe is required to turn the new pulley in brass or aluminium.

To get at the drive mechanism one has to proceed with the following steps after the 8555A has been withdrawn from the 141T mainframe:

1. Remove the top cover (6 screws).
2. Turn the frequency controls to lowest frequency (fully anticlockwise) and ensure that the control is in the "normal" tune position (i.e. pushed fully in).
3. Remove the frequency control knobs via the small hex screws.
4. Next, with plug-in front panel facing you, unscrew the two countersunk crosshead screws on the left hand side of the unit and nearest the front panel.
5. Invert the unit and unscrew the two countersunk cross-head screws on the bottom of the unit nearest the front panel.
6. Looking again on the top of the unit, locate and carefully prise off the D type connector just behind the display. (This has a large loom of wire going to it). The whole front display including front panel should now lift upwards and away form the rest of the unit.
7. Remove the front panel escutcheon by using an Allen key to undo the "Ext mixer bias" and "Amplitude cal" pots. and also remove the circlip around the frequency control spindle.
8. This exposes three crosshead screws underneath the "BAND" escutcheon which should be removed.
9. At the rear of the front panel in amongst the front panel switches are two large nuts. One only has to loosen them off one turn, since the drum assembly is held in position with slotted brackets (clever HP!). The whole tuning drum assembly will then lift clear of the front panel.
10. Locate the bracket holding the three blue multi-turn pots and very carefully remove it by taking out the two cross-head screws holding the bracket to the display. Do not disturb the rotational position of the pots. The long tuning spindle will also come with this assembly together with the "normal / rapid" tuning clutch assembly. This contains 4 sprung ball bearings so be very careful with this!
11. This will leave on the display section the brass bracket carrying the display drive sprocket and a gear.
12. Remove this brass bracket by taking out the two long cross-head screws and then remove the circlip which holds the sprocket on the hollow shaft. You should now have complete access to the sprocket.

A new pulley can now be turned up in brass or aluminium, with a diameter of 19.2mm, a lip diameter of 20.3 mm and thickness of 10mm. as per the diagram. A hole with diameter



New Pulley Detail

of 9.5mm with a small key in it is also required to key it to the keyway in the existing shaft.

This can be made by filing a small slot in the pulley and soldering a small piece of brass in the slot.

The old sprocket can be used as a pattern. The new pulley must be a good push fit on the shaft otherwise backlash and wobble will occur.

For those unable to get access to a lathe, the original sprocket could be modified by carefully filing off the sprocket teeth and sticking two, slightly oversized, "cheeks" in thin brass or aluminium to the sprocket, thus effectively changing it to a pulley.

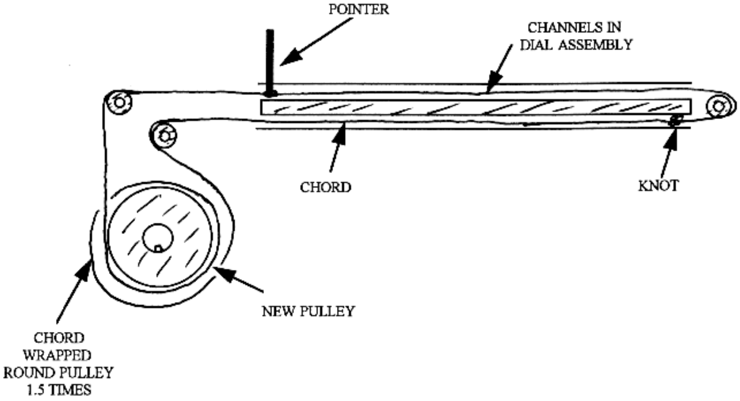
To re-assemble replace the hollow shaft including the new pulley and fasten with the circlip into the brass bracket and screw the bracket back on to the dial assembly.

Since the chord used for dial drives seem to be obsolete one now has to improvise! Fishing line of 2-3mm diameter, or something very similar, can be used as long as it will not stretch with time. This is threaded over the path shown in the accompanying diagram and wound round the new pulley 1.5 times. Care should be taken not to loose the 3 small pulleys over which the chord runs. The tricky part is getting the line knotted and in good tension at the same time.

Three hands can be useful!

Note the knot should run in the lower slot and must not be so large that it prevents the chord running freely. The purist might like to add a small spring at this point to maintain tension but, in practise, the loading on the chord is so light that this has not been found necessary. Ensure that the knot is located at the right hand end of the slot as shown and reverse steps 8 to 13 above, taking care again not to move the pot. shafts.

Rescue the aluminium pointer from the old belt by carefully cutting it off, but leaving enough material to fix it to the new chord with a blob of Evo-Stick. A small white mark exists on the tuning dial just above the words above "L.O." and the pointer is located opposite this mark. Ensure again that the tuning shafts are in the fully anticlockwise position i.e. lowest frequency during this process. Before full re-assembly check the free running of the chord from the low to high frequency end of the scale. Reverse steps 1 to 7.


POINTER
CHANNELS IN DIAL ASSEMBLY
CHORD
KNOT
NEW PULLEY
CHORD WRAPPED ROUND PULLEY 1.5 TIMES

The unit can then be placed in the main frame and checked for frequency calibration against known signals. Providing the tuning pots have not been moved during the process and the pointer has been fixed in the position shown, calibration should be as per the original.

The HP432A Microwave Power Meter
A legend in its own lifetime

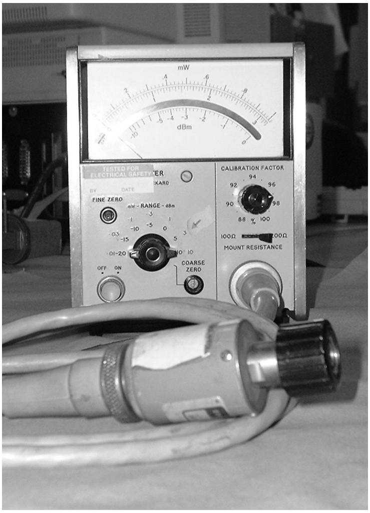
John A Share

Introduction

There can be few items of Instrumentation that have been in production for over thirty years. The HP432A Power Meter was designed in the late 1960s, entered production in the early 1970s and will cease to be available after September 2003. Entirely discrete analogue design, devoid of a digital display and built onto antique style circuit boards, these units continue to be used in numerous situations and applications throughout the world. They often appear on the second user market at a fraction of their original cost (currently c£10,000) and are a tempting purchase.

Model options relate to the location of the thermistor input connector and the provision of the battery pack. Option 001 has the rechargeable battery pack, option 002 has a thermistor input on the rear panel wired in parallel with the connector on the front panel whilst option 003 has the thermistor input only on the real panel.

Thermistor cables are available in 10ft, 20ft, 50ft,100ft and 200ft lengths and are suitable for use with any type of thermistor be it 100 ohm or 200ohm. The thermistors themselves are available in Coaxial or Waveguide Mounts and have a frequency range of 0.01 to 40 GHz. The most common heads are the Coaxial "N" type. They resemble a pepper pot, and whilst HP list only two models other manufacturers have produced alternatives. Most second user HP432As seem to be equipped with the 0.01 to 10 GHz head (HP478) and these are relatively inexpensive. The 0.01 to 18 GHz head (HP8478) is rather scarce and usually priced at a premium. Alternative 8 GHz heads are plentiful and are often priced at a few pounds each.



A calibration table should be found on the side of the head. With an HP unit this takes the form of an aluminium label with calibration points marked with a centre punch. For an accurate power reading at a specific frequency the HP432A calibration switch is set to the value marked on the label. Unless this requirement is critical such precision can be ignored and a median calibration setting used for the entire frequency range.

Coaxial heads have an internal resistance of 200 ohms. More exotic heads designed for direct connection to wave guide have an internal resistance of 100 ohms hence the two range options on the HP432 front panel.

The Package

For the Power Meter to be regarded as a usable system it needs to consist of three items, the meter itself, a head cable and a thermistor head. The battery is not essential, it is optional, and an Option 001 meter will function without this item. Head Cables are quite specific and, whilst it would pose few problems to make a suitable four core cable, locating a source of the connectors poses a serious difficulty. Fortunately these cables are available on the surplus market for a few pounds each. Low frequency thermistor heads (e.g. < 8GHz) are also readily available but intending purchasers are warned against buying heads that have been "repaired" or show obvious signs of tampering. They are notoriously difficult to repair. When purchased from a professional second user source, the package should include the workshop manual. These same sources are sometimes willing to make photocopies or supply originals but at a significant price. Without the manual these units are impossible to repair.

Circuitry

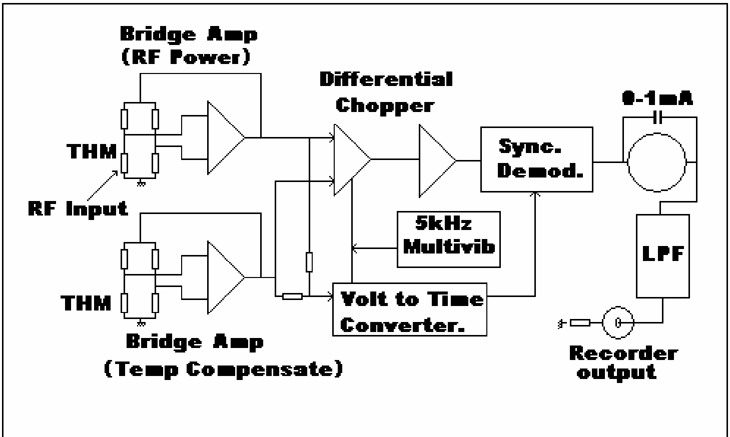
Input power warms the RF Power Thermistor creating a voltage (V_{rf}) proportional to the temperature change of the Thermistor, room temperature is sensed in an identical manner to create a compensation voltage (V_{comp}). The change due to the RF input is therefore the difference between these two voltages ($V_{rf} - V_{comp}$). Amplifying small dc signals poses serious base line stability problems and the classical solution was to convert the dc signal to ac, amplify, and then rectify back to dc. The term for this technique was a "Chopper Amplifier". Today there are Integrated Circuits that perform this role, e.g. 7650, and certain Operational Amplifiers are available to compete with their specification. In the HP432A this is performed using discrete components. Unfortunately ($V_{rf}—V_{comp}$) does not provide a "Power" measurement, it is necessary to include an "R" term as in the basic equation $P=V^2/R$, it is also necessary to remove the V_{comp} term. To achieve this the two values are summed ($V_{rf} + V_{comp}$) and the resultant value modifies the width of the 5 kHz signal to the synchronous demodulator. The demodulator output is therefore a pulse train of height proportional to the difference, and width proportional to the sum of V_{rf} and V_{comp} . The capacitor across the meter is an essential part of the integrating circuit that causes the meter to display $(V_{comp})^2 – (V_{rf})^2$. The Low Pass Filter is an integral part of an Auto Zero Circuit and the user output "Recorder Output" is a dc level devoid of the pulse train into the meter circuit.

Maintenance

Experience with a number of HP432A over the past fifteen years has clearly demonstrated that they are very reliable. There has never been a catastrophic failure and they are extremely easy to repair.

Whilst HP insist that of the sixty semiconductors used in the design over half are selected devices, it has been found that they can be generally replaced with off the shelf items of similar specification and some quite simple device testing to form matched pairs. One or two differential pairs are exotic devices and might prove difficult to locate they can, with a little ingenuity, be substituted.

The majority of failures have been due to ageing of switch contacts and judicious application of switch cleaner has affected a ready cure. A single case of a "sticky meter" was traced to the meter scale flaking and depositing flecks of white paint into the move-



Simplified Outline of HP432A Power Meter.

ment. Despite numerous attempts to halt the decay the problem continued. Eventually the meter scale was photocopied, the original paint removed entirely and the photocopy glued in place ... an invisible repair! Damaged heads must be regarded as irrepairable.

Applications
The HP432A has seven ranges, the most sensitive being 10 microwatts. The open end of a Coaxial Thermistor mount is an N type male connector. Numerous attachments can be fashioned onto N type sockets so that the HP432A functions as a very sensitive signal detector. In this manner wave guide modes have been determined by fashioning a dipole and rotating it to detect TE or TM modes within a guide that was excited from a signal generator. A closed loop will function as a general "sniffer" for detecting signals in a multiplicity of circuits. Low power circuits, for example amplifiers, can be connected directly to the thermistor head; it then functions in a dual role as load and detector.

For measuring real power, the least sensitive range of 10 milliwatts might appear somewhat limiting until it is realised that Coaxial Attenuators are plentiful and very inexpensive. Typically -10db, -20db or -40db with frequency ranges up to 18 GHz, when used in conjunction with directional couplers it is possible to measure immense amounts of power. In one application we use a -30 db bidirectional coupler / -30db attenuators / HP478 heads to measure the forward and reflected power in a system using a Klystron Amplifier that delivers 750 watts at 8.2 GHz.

Integrating an HP432A into a computerised system is made difficult because of the absence of remote range selection. In reality, this has not been found to be a real obstacle. However, the inherent drift that requires repeated "Zero Resets" can be automated. The reset circuit is simply a biased toggle switch and a pair of relay contacts wired directly in parallel with the switch and driven from PC interface card is a simple solution. The switch is readily accessible and there is adequate space to mount a small relay on a custom PCB within the enclosure.

On the rear panel of the HP432A is

a "Recorder" output. This has a maximum amplitude of 1.000 volts for full scale deflection of the panel meter, irrespective of sensitivity range. Connecting this to the input of an ADC within a PC through a long lead is prone to problems, mostly related to earth loop induced 50 Hz. The Recorder shield is not grounded at the panel; it is bushed. Ideally the connection to this should be differential but the connector is not two pin! Similarly digitising a remote single ended 1 volt signal with a 10 bit ADC invites problems of noise and hence loss of resolution. A working solution has been to pass the HP432A output through a passive 50 Hz notch filter, amplify the signal using a differential chopper amplifier to 10 volts and connect this signal to the single ended ADC input through the long lead. The difference is impressive!

There is a delay between the application of the input signal to the head and the moment when the output is a true representation. This is not fixed but depends on the step amplitude of the input signal. For the Analogue display this is of little importance. However, when using a PC, it is all too easy to forget that there is a significant delay (in PC processing terms). In general a full range step input change of 10mW requires 1 second of delay but this can be reduced to 100 ms with virtually no loss of accuracy. Where the input is fluctuating it is possible to reduce this delay even further but experience has shown that 20ms is probably the lower limit and the thermal lag of the head then becomes noticeable.

Conclusion

Many of my HP432As are well over twenty years old. They have proven to be incredibly reliable and have been used in many, many roles and applications. They are simple to configure, easy to use and give unambiguous readings. Down time figures would indicate that they have been by far the most reliable items of equipment I have ever purchased.

The new price of an HP432A is daunting; their availability on the second user market at one hundredth of that price is certainly attractive and, when seen at surplus sales at one thousandth of that original price, they represent the steal of all time.

Repairing HP Power meter heads

Tom Williams, WA1MBA

Several people have asked me if it is possible to repair an HP 478A-style thermistor mount (Power Head for 431 / 432 power meters). Surprisingly, in SOME cases, the answer is YES!

The first step is to do triage - figure out if the head is completely fried. Some of these heads will withstand 1 watt for 10ms and appear broken but can be brought back to life.

The only way to tell if a thermistor resistance is within specification is to measure the resistance under simulated operating conditions. This is complicated and I'll outline how to do it at the end of these notes. In the case of a power head that pins the power meter (one way or the other) you don't care about resistance being within spec - IT ISN'T. What you do care about is whether it is totally fried or can be brought back.

To determine if it is repairable, measure the resistance between pins 1 and 2, and also between 3 and 4. The resistance should measure between 1000 and 5000 ohms. Use a good digital ohmmeter on the 20K scale so that it draws very little current. Don't leave the meter connected for more that a second or two. The absolute maximum current that a thermistor can take is 13 mA. If you read an open or a short, you have blown thermistors, and might as well trash the unit. If they are both about 1K to 5K, proceed to repair/ adjust.

Don't forget to check the cable... if there is an open, or a faulty connection, this can cause the same symptoms as a bad head.

The procedure is as follows

1. Take the can off the N connector by removing three set screws. Slide the cover can off.

2. Plug the unit into your 431 / 432.

3. Set the resistor mount ohms to 200 (assuming you are repairing a std 478A - some of the microwave heads are 100 ohms, so set the control accordingly).

4. Turn the power on.

5. If the meter is pegged down scale

a. Set Range to 10mW

b. Set ZERO and VERNIER to mid-range (on the 432, carefully count the turns of the screw-driver coarse- zero pot and set it to mid range)

c. On the back of the thermistor mount there are two small brass screws.

Take your time. Turn one of the screws 1/8 (yes 1/8) turn clockwise. Then turn the other screw 1/8 t clockwise.

If there is a sudden jump in meter indication when advancing either screw, back it off 1/8 turn, and do NOT advance that screw any further. If either screw bottoms, do not apply force - it is likely that if a screw bottoms, the thermistor is fried.

d. The best result is when at some point in the alternating "turning of the screws" the meter rises.

Once it starts to rise, trim it to zero by turning each screw a little.

6. If the meter is pegged upscale,

a. Set meter to Zero (as in 5 above step b)

b. Set RANGE to highest power position which will not peg the meter.

c. Turn one of the little brass screws counter clockwise (leftwards) to obtain a meter reading of half the deflection noted in step b.

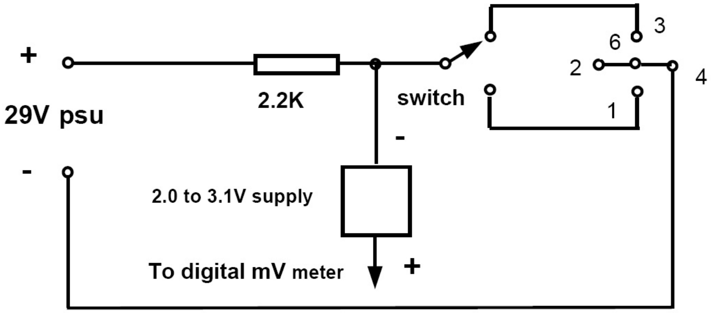
d. Turn the other screw counter clockwise to zero the meter. If it is impossible to zero the meter, at least one of the thermistors is fried.

7. Replace the cover.

Congratulations! The head works, is no longer calibrated, but is probably within one dB of its original calibration. Check it with a known good source or another meter to get a calibration factor.

Note. Some waveguide type units have only one adjusting screw. Follow approximately the same procedure, adjusting only the one screw.

If you really want to check a mount for thermistor match you need a 29V DC power supply, a second high resolution supply with floating terminals adjustable 2 to 3.1 V, a switch, a 2200 Ohm resistor and a millivolt resolution digital volt meter. You set up the circuit below and adjust the adjustable supply as a bucking voltage to get a very small reading on the volt meter and then switch between the two thermistors to make sure that the readings do not vary by more than .030 volts. Non-operating units with readings as high as .150 difference can usually be repaired per instructions. I don't recommend that you do this, first of all, its not usually necessary, and second of all, if you get it wrong you can fry the thermistors!



Curing Cyclic Instability and other problems in the Adret 5104 synthesizer

John Hazell, G8ACE

This Adret Synthesiser, new to me, had fully working frequency switches and the reference oscillator was set correctly.

However, the red alarm light had a cyclic flash and the output would jump just a few Hertz.

The Adret is not the easiest unit to work on without extender boards. With the initial investigation limited to checking signals on the linking cables, a blank was drawn as the fault appeared to be everywhere. By substitution, the fault was narrowed down to 6883, the 1MHz Spectrum Module. It was found that the PLL controlling the 50MHz oscillator was hard at one end, the varicap volts being at around 11.5v. The 50MHz is locked to the 10MHz reference so this was duly cranked and the PLL came off its stop with the reference 150Hz low. So the loop was capable of working and the instability disappeared. A 'Quick' fix is always desirable to prevent the unit joining those on the 'too difficult' shelf! A 15pF capacitor was inserted in series with the crystal to raise its frequency. The crystal, it is assumed, must have aged somewhat.

The value of the C is quite critical. 15pF centred the loop at 6.8v. The 10MHz reference could then be varied -100Hz/ +50Hz with the loop remaining in lock.

This was considered an adequate fix.

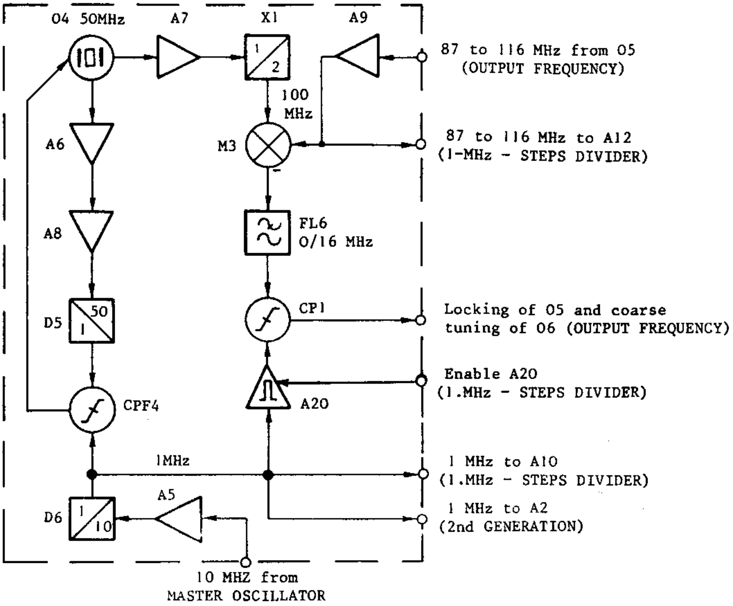
A quick check for the 50MHz loop, therefore, is to vary the reference oscillator over its whole range whilst viewing the Hz digits of the output on a counter. Do this only if you can reset the reference correctly of course. The block diagram of the 6883 module is shown overleaf, along with the circuit for the 50MHz oscillator taken from the Adret manual **which is essential if one is to service this unit ... editor**.

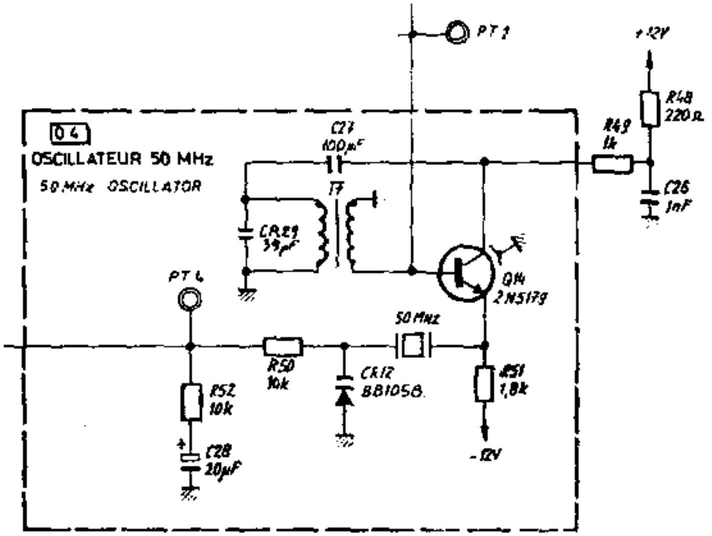
Another stock fault is the prescaler chip, a Plessey device. I replaced it with two 10131 chips wired as divide by 10 to cure the following fault. The front switches fail to work, the output being stuck up around 130MHz. Chris, G8BKE has had and fixed this fault recently. I have had it and so has another local, so it’s fairly common.

The other little snippet is the genlock input.

It’s always active waiting for a 5MHz signal. If the internal oscillator is, in my case, within 20Hz, the oscillator simply snaps in. Another unit, newer, needs to be within 4Hz.

There’s no need to touch the master frequency knob, unless of course you want it to be right without the external signal.


O4 50MHz
A7
X1
A9
87 to 116 MHz from O5
(OUTPUT FREQUENCY)
100 MHz
M3
87 to 116 MHz to A12
(1-MHz - STEPS DIVIDER)
A6
A8
FL6
0/16 MHz
D5
50
1
CP1
Locking of O5 and coarse
tuning of O6 (OUTPUT FREQUENCY)
Enable A20
(1.MHz - STEPS DIVIDER)
CPF4
A20
1MHz
1 MHz to A10
(1.MHz - STEPS DIVIDER)
A5
D6
1
10
1 MHz to A2
(2nd GENERATION)
10 MHZ from
MASTER OSCILLATOR


PT 1
+12V
R48
220Ω
R49
1k
C26
1nF
O4
OSCILLATEUR 50 MHz
50 MHz OSCILLATOR
C27
100pF
T7
CR21
39pF
Q14
2N5179
PT 4
50 MHz
R50
10k
CR12
BB105B
R51
1,8k
-12V
R52
10k
C28
20µF

More Notes on the Adret 5104 Synthesiser

Mike Scott, G3LYP

A while back, I bought an Adret 5104 which, on testing, I found to be faulty, giving output on about 125MHz which was not controlled by the switches on the front panel. I then obtained a second unit which was working, and by swapping plug in modules between the two units, I found that the faulty module in the first unit was the output module designated 6887.

At the time I had a quick look inside the module and noticed that some of the PCB was a bit brown, indicating that it had been quite hot. As the second unit was working I decided to keep the first one as spares.

I wanted to make a signal source for testing AO-40 antennas and receivers using the Adret and a modified DK004 LO source. Unfortunately when I switched on the Adret, I found that it was no longer working, giving output on about 74MHz which, like the first unit, was not affected by the switches. As I had previously established that all the modules in the first unit, with the exception of the output, were working, I thought that there was a good chance that the fault in the second unit would be in a different module. Not so! The output module was again the faulty one.

The Adret is a very complex device and after going through the manual several times, I still don’t fully understand it. However, there was no choice but to have a go!

One of the problems in servicing the units is that, because of the modular design, most of the circuitry is in metal boxes plugged into a motherboard and it is virtually impossible to work on the modules while they are in situ. The manual refers to extender cards which are used to make the modules accessible while remaining connected to the motherboard.

The 6887 output module has only an eleven pin connector and not all of the pins are used (some of the modules have 35 pin connectors) so in my case it was possible to solder wires to the mother board and to the PCB and so get access to the module on the bench while still connected to the main unit. There are also a number of SMB connectors on the top of most of the modules and by using a commercial “T” adapter or fabricating one from bits of coax and SMB connectors, it is possible to monitor the signals at many points in the unit. The manual recommends the use of a spectrum analyser and 100MHz oscilloscope, neither of which were available to me. Nevertheless, with a frequency counter, a multimeter, and a 20MHz ‘scope, it was possible to do quite a lot.

The Manual contains several fault trees, but working through these confirmed what I already knew, that the fault(s) were in the output modules. These modules contain three basic circuits.

The first is an oscillator which covers 90 – 120MHz, and which provides the output of the Adret after buffering and amplification.

The second circuit is a virtually identical oscillator covering 87 – 117MHz which is fed back to the 1MHz Spectrum module (6883).

Both oscillators also provide output, via separate buffers to a mixer (an MCL device similar in appearance and pinout to the SBL- 1).

The 3MHz IF from the mixer is

passed through a multistage filter to the 11 pin connector and via the motherboard to the Phase/Frequency module (6886). There are also a number of other bits of circuitry associated with the control of the two oscillators. On the first of the faulty modules, a check of the signals at the various SMB connectors indicated that there was no output from the 87 – 117MHz oscillator sub-unit. A check with a multimeter revealed that the 2N5179 base-emitter junction in the buffer between the oscillator and the output socket was defective. This transistor was replaced with a BFY90 which is listed as an equivalent. After replacing the module in the motherboard, the unit started to work normally.

Unfortunately, tracing the fault in the second unit took considerably longer. I established that both oscillators were working, although the output of the 90 – 120MHz oscillator was at about 75MHz. The 87 – 117MHz oscillator appeared to be working normally and was controlled by the switches on the front panel. On checking carefully, I discovered that it worked up to 114MHz but, on switching the 1MHz switch to the next step (115MHz), the frequency fell back to 107MHz. On checking the DC level on the "Approach" line to this oscillator I found that the voltage at Test point 1 (All the five test points on this module were connected to colour coded wires and brought to the top of the module to enable monitoring while the module was in situ) rose steadily to a similar level as on the now working module (approx. 8v) but on switching to the next step, the voltage fell back to about 6.8v. On the good module the level rose to 8.8v. The "Approach" line is fed via a 4.7k resistor (R9 to a group of four transistors (Q8 – Q11) which compare the "Approach" signals from both oscillators and feed a signal back to the Phase/Frequency Comparator. On the working module the voltage at Test point 3 (the bases of Q8/Q9) was only marginally lower than at TP1 and ranged from 3v to 8.8v depending on the switch settings. On the faulty unit the voltage at TP3 was very low (less than 1v) on all switch settings. The immediate conclusion was that one of the transistors in this group was faulty. Unfortunately, checking with a multimeter suggested that all were in good condition, but as the voltage from the meter was only about 3v, I decided to replace all four. Luckily, all were common types and easily replaced. This was a good decision, because on re-installing the module, all was well and the output ranged from 90 – 120MHz with all the switches functioning as expected.

Having a working unit for comparison was a great help, and it would be useful if the Microwave Committee could keep a record of Adret owners who might be willing to help owners of faulty units.

As a final note, not having the correct mains lead, I replaced the mains connector on the back of both my units with IEC connectors fitted with integral filters.

The MACOM PE334 Synthesiser

Andy Talbot, G4JNT

At rallies and Microwave Roundtables there have been for sale a range of microwave synthesiser bricks.
The data sheet supplied with these had several errors and this note has been provided to help users get them working.

Power supply requirements:
+20V at approximately 400mA for the analogue circuitry on the pin marked
+5V at 150mA for the logic

Frequency control is via a 26 way IDC type connector, with TTL level signals on 12 parallel lines. Logic level is positive true, and the lines float high to a logic '1', they need to be pulled to ground for a logic '0'.

The pin configuration is detailed below, shown looking at the synthesiser module.

There are four types of synthesiser module covering different frequency bands and step size:

SAC12-01
4.7 - 5.1GHz Step size 0.625MHz

Pin 1 is missing.
25 23 21 19 17 15 13 11 9 7 5 3 (1)
26 24 22 20 18 16 14 12 10 8 6 4 2

SAC32-00
6.35 - 6.76 GHz Step size 0.25 MHz

Connections :

1 to 11	No connection	19	B5
12	B11 (MSB)	20	B4
13	B10	21	B3
14	B9	22	B2
15	B8	23	B1
16	B7	24	B0(LSB)
17	B6	25	Ground
18	No Connection	26	Ground

SAC32-01
6.65 - 7.16 GHz Step size 0.25 MHz

SAC42-01
10.95 -11.45 GHz Step size 3.125 MHz

Programming
The three lowest frequency units are programmed by applying a straightforward binary code to the 12 parallel programming lines.

The frequency of each of these can be set by determining the value of the programming number, N, then applying this value to the lines.

The 11GHz module requires two separate six bit programming numbers to be calculated.

Got to the next page ….

SAC12-01 Freq (GHz) = 2560 + 0.625.N or N = 1.6 * (F - 2560)
SAC32-00 Freq = 6160 + 0.25.N or N = 4 * (F - 6160)
SAC32-01 Freq = 6144 + 0.25.N or N = 4 * (F - 6144)
SAC42-01 Freq = 8000 + 125.L + 3.125.M

For example, to set the SAC32-00 to 6400.25MHz,
N = (6400.25 - 6160) * 4 = 961
Convert to Hex / Binary = 0x3C1 = 001111 000001

Only frequencies that are within (or close to) the allowed frequency bands can be programmed, thus the full range of programming numbers possible are not all valid, any attempt to set these invalid codes results in the synthesiser VCO free running. It is possible that by adjusting the VCO centre frequency by tuning the cavity, the lock range can be moved. This has not yet been tried.
The X band module, SAC42-01 is more complex in its frequency setting than the other types. The control lines do not all follow a simple binary sequence but are weighted as follows :

B0 - B5 Six bit binary value M giving 64 * 3.125 MHz steps up to 200MHz
B6 - B11 Six bit binary value L giving the 125 MHz steps.

Note that in some cases a specific frequency may be obtainable with more than one programming code, codes within the ranges shown in the examples below have been tested :

10850 MHz = 0x5A0 = 010110 100000 Below lower freq spec.
11000 MHz = 0x600 = 011000 000000
11125 MHz = 0x628 = 011000 101000 Two codes for this freq
11125 MHz = 0x640 = 011001 000000 " " "
11275 MHz = 0x670 = 011001 110000
11400 MHz = 0x6B0 = 011010 110000
11571.875 MHz = 0x6FF = 011011 111111 Above frequency spec.

Ailtech Noise Sources

Charles S. Osborne, K4CSO

At the risk of losing some choice hamfest finds, a posting of an Ailtech noise source on EBay reminded me of something I've discovered and should pass along about those noise sources:

If you have an Ailtech 7616 noise source or variant that won't calibrate, i.e. seems dead, don't give up on it. The noise source may actually be a separate module inside, made by Noise Com, MSC, or unlabeled others. Usually what goes is the attenuator which is used to calibrate the unit. It's a thick film hybrid. Over tightening the N-connector can twist the connection, cracking the attenuator's ceramic substrate, open circuiting it ... Or, at very least, it will be very unpredictably intermittent.

Now the good news ... most of the units I've disassembled had SMA connectors on the noise source internally and a normal SMA inline attenuator following. Disassemble the unit down to the noise source and use it with a new SMA inline attenuator and you are often times back in business.

I've bought a number that looked like they had been drug up and down the flight line, if not outright submerged. Most were in the $5—$10 range since the labels were gone and they were clearly a long shot. I'm 4 for 5 working once I tossed the cracked or fried 16dB input attenuator.

The noise source is typically 250,000 K or 28-30dB ENR without the 16 dB attenuator. The attenuator value is factory selected to get the unit into the 15.4 dB ENR = 10,000 K noise customary range. Changing to a good 18GHz 26dB attenuator (even if its a 16+10dB arrangement) gives you a 5dB ENR unit which is better for measuring extremely low noise preamps (if your noise figure meter will calibrate at 5dB ENR).

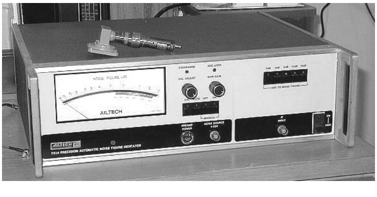
If you go to a conference where a noise figure meter is available, you can measure the same preamp with a known head and compare your "rebuilt" one and get even closer on the ENR re-calibration at that spot frequency.

I guess my secret's out ... no more $5 noise sources once this group gets turned loose on them at the hamfests! Now, if I could just revive my 20 year old HP8970A to use and calibrate the ones I have.

Agilent says they obsoleted the HP8970A in 1987 and its successor the B version in 2001. Story of our lives, right?

The interesting thing is they are offering $10K to get them out of our hands in a buy back deal on one of the new 8974A $40K noise figure analyzers. I've got about as much chance of getting one of those as a 100ft yacht.

Below: The Ailtech 7514 **P**recision **A**utomatic **N**oise **F**igure **I**ndicator (PANFI)
This item is, from time to time, found on the UK surplus market and is well worth acquiring if you can also find or make a noise head.



Care and feeding of Minicircuits Lab ZHL42 amplifiers

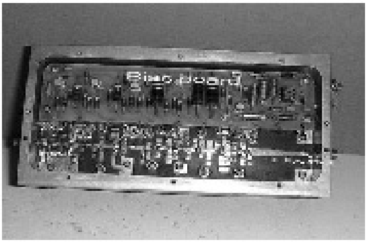
Sam Jewell, G4DDK

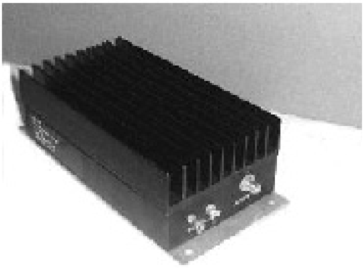
Minicircuits Lab introduced these amplifiers in the late 1980s as high power, linear gain blocks for commercial and industrial lab use. They were quite expensive and used then new GaAs technology.

A wide range of variants have since appeared including 10MHz to 4.2GHz types (ZHL-42W), 40dB gain types (ZHL-4240) and lower gain 100mW output amplifiers (ZHL- 1042J). However, the GaAs FET input stage was very prone to being destroyed by too much signal. A very common failure situation was simply turning on signal generators that didn't remember the last set level and that would power up with maximum output of perhaps 50mW.

This was almost certain to cause the amplifier to fail. When they first appeared they were rated at +10dBm maximum input, this was later revised down to +5dBm and my own advice is to ensure the input doesn't exceed 0dBm.

The purpose of this article is to show that these amplifiers can easily be made to work again and are (arguably) more useful to the amateur community after this simple repair than in their 'as new' condition. Why? How? Read on!

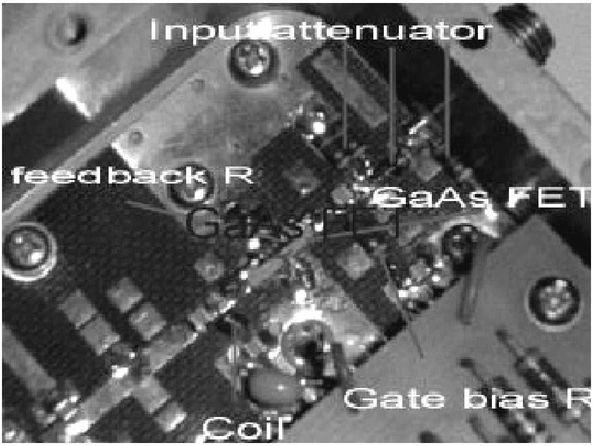




700MHz to 4.2GHz 1Watt output, 30dB gain, 1dB ripple broadband

After trying unsuccessfully to replace the blown input FET I tried the more practical approach of merely strapping out the input stage. First, check this is where the fault is. (You can be better than 90% certain it will be). It is easy to prove this is the stage at fault as the bias condition around the device will be found to be inconsistent with, usually, the negative gate bias voltage on the first stage non existent. Carefully remove the 820R gate bias feed resistor, 680R feedback resistor together with series chip capacitor, close-wound drain feed inductor and then the GaAs FET device. Cut a 1mm wide, 12mm long strip of thin copper tape to bridge between the gate and drain connection points. That's it!

With this 'repair' the gain of the amplifier will be found to have decreased by about 10dB. However, it is now possible to increase the input signal level significantly without fear of damaging the amplifier. I have applied up to 100mW without damage after making these changes. The great thing about this mod is that the achievable



output power from the amplifier is now significantly increased.

The following table shows the results from one such amplifier 'repaired' in this way. You will note that the frequency response is now noticeably less flat than 1dB. However, for amateur purposes this probably doesn't matter too much. Some additional gain can be obtained by removing the input 3dB attenuator and replacing it with another copper strap.

These amplifiers are an ideal way to make a multi-band transverter or to boost the output of your PLL Brick oscillator to produce a useful output power for a 23 - 9cm, personal beacon.

Frequency	**432MHz**	**1.3GHz**	**2.32GHz**	**3.4GHz**
Output power (sat)	1.4W	2.25W	1.7W	2.0W
For input power	Not measured	10mW	10mW	25mW
700mA at 15V	Rising to 875mA at saturated output power			

Finding a Use for Defective Marconi Power Meter Heads

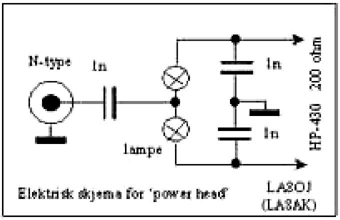
Jan-Martin, LA8AK

The following information was gleaned from the website of the late Jan-Martin Noeding, LA8AK.
It's presented here as a starting point for what could be the "answer to a maiden's prayer" as far as those of us who have blown Marconi power meter heads in the bench drawer.

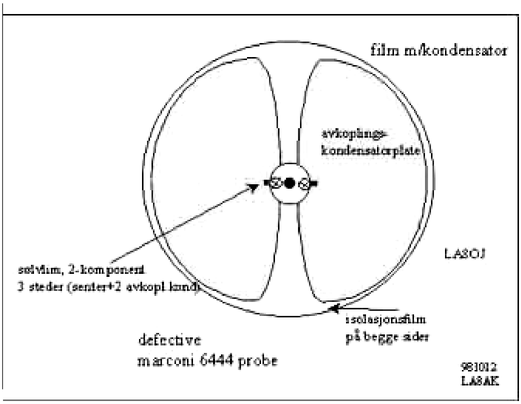
Although the diagrams are not very sharp, as taken from Jan-Martin's webpage, they show the general idea of using miniature watch illumination lamps to make a workable sensor.

The editor would be very interested to hear from anyone trying this out and also from someone with a source of the miniature lamps. A check at the Conrad website, the company mentioned in the diagram below, failed to find anything suitable.

Using defective MI 6444 probe to repair old HP 431B and Narda power meters



Repairing power head using miniature watch illumination lamps from Conrad in Germany
As shown for Marconi power head


film m/kondensator
avkoplings-kondensatorplate
LA8OJ
sølvlim, 2-komponent
3 steder (senter+2 avkopl.kond)
isolasjonsfilm på begge sider
defective marconi 6444 probe
981012
LA8AK

Chapter 15

Mechanical

A Tilt-Over Mast for Portable Microwave Operation

Jean-Paul, F5AYE

(English translation by G3PHO)

The System offers two important functions:
1) Easy assembly of antennas, with the mast in the horizontal position and then erection of antennas and mast with minimum effort.
2) Perfect verticality, even on uneven ground

An adjustable, ladder-like structure was constructed and fixed horizontally to make the upper support frame for the system (see figs. 1 and 2 below). It is bolted to solid metal roof bars on the vehicle. The mast pivots at the height of the vehicle's tow bar ball joint. (see figs 6 and 7).
The whole assembly can be dismantled after portable operations are over.

Some comments:

- The higher the roof of the vehicle, the more rigid the mast structure.
- If the roof bar supports are high you can stiffen them further by cross-bracing from one to the other (see fig. 3).
- The description is given here without sizes and dimensions, but it should allow anyone to visualise their own version with its own possibilities and materials.
- In my case I used a lot of aluminium. The mast is telescopic and, when wound down, can stay in

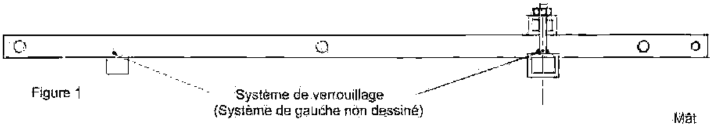


Figure 1

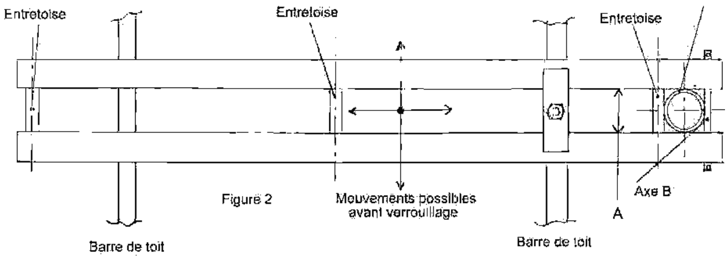


Figure 2

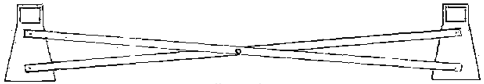


Figure 3

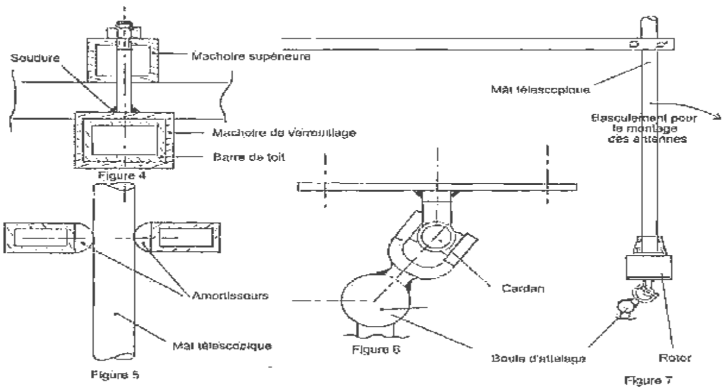


Figure 4

Figure 5

Figure 6

Figure 7

position during the journey to and from the site.

- The lower joint is a home made item (see fig 6) but a reclaimed car universal joint or similar would serve just as well.
- The easiest way to install this joint is to weld it to the ball joint but there are certainly better, alternative methods.

Specific Points:

- Removable spindle or axle (B) is held in place by two pins.
- Leave sufficient space between spindle B and the support frame cross beam so that it allows the mast to be tilted up and down between the two sides of the ladder structure.
- Cover the cross beam, where it contacts the mast, with some kind of synthetic material such as hose pipe off cuts so as to reduce vibration noise during the journey to and from the portable site.
- On the ladder structure side arms, fix 2 half round pieces of synthetic material to form
- shock absorbers which can cut down noise and, above all, allow a high mast height with the minimum of play (fig.5).

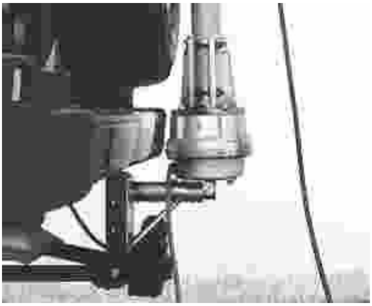
In Use:

- Insert the mast, without antennas, into the end section of the ladder and retain in place by inserting and pinning axle B (see photo lower right)
- Slacken the clamps (fig. 4) so that the ladder structure can slide backwards and sideways on the roof bars.
- With the aid of a spirit level, set the mast in the vertical position.
- Tighten the clamps holding the support frame to the roof rack.
- Remove axle B in order to tilt the mast into the horizontal position for mounting the Antennas

EXPLANATION OF FRENCH TERMS USED IN THE DIAGRAMS:

VERROUILLAGE:
Bolt or clamping system
ENTRETOISE: Cross members
BARRE DE TOIT: Roof bar
MAT: mast
AXE: Axle or swivel bar
AMORTISSEUR: Shock absorber
BASCULEMENT: tilt-over/lowering
SOUDURE: welded joint

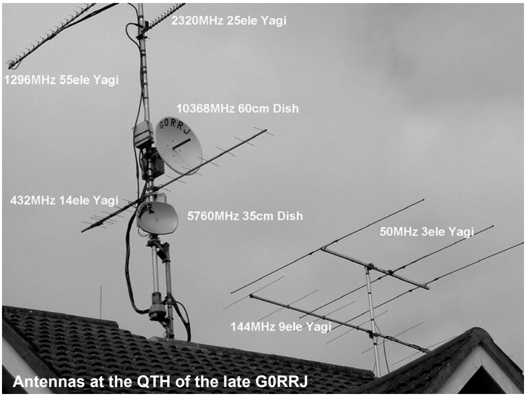
Below left:
The rotator stands on a home made bracket



Below right:
The Upper support holds the mast firmly

Will Your Antenna Stay Up Next Winter?

Mat Reilly



Antennas at the QTH of the late G0RRJ

From WA1MBA's microwave reflector on the Internet… it should be of interest to those readers who have or who are thinking of having a home station microwave antenna based on the old principle: " If it stayed up this winter it wasn't big enough" !

This is a gross rule of thumb. It derives from the Rad Lab series. Back during WWII, the folks at the Rad Lab did a study on dish wind resistance ... lots of wind-tunnel models, studies, measurements but all relatively crude. They came up with 40lb/sqft of antenna area in a 100mph wind. Andrew re-did all the studies in the sixties with much more expensive wind tunnels, lots of different orientations of the dish in relation to the wind direction, lots of different materials/surfaces/ meshes. They came up with a max of about 40lb/sqft of antenna area in a 100mph wind.) So, the worst case force on the dish will be

F = k*v^2

This is a gross over simplification but I'm not sure this is the place for all the equations that none of us really understand.

Gentle reader, if you understand the governing equations, especially under conditions of compressibility, please send me mail, as I'm looking for some advice on numerical solutions to the Navier-stokes equations.
The empirical evidence is that

k = 40 lb/ft^2/10^4mph^2

So, a 30" dish in a 50mph wind produces:

F = (40 / 10^4) * 50 * 50 * 1.25 * 1.25 * 3.14
= 50 lbs.

I wish to heck I could come up with the reference! When I'm in the files next, I'll look for it. The major interesting tid-bit I got out of the Andrew study was that the maximal force was, of course, NOT when the wind was normal to the face of the dish.

As every sailor knows, you get a lot more force out of the sail when it is acting like a wing than you do when it is acting like a bag.

Self Locking Screws for Microwave Adjustments

André Jamet, F9HX

We often use screws to adjust microwave devices like filters made with copper, plumbing caps, frequency adjustment of DROs or as matching sections in a wave guide.

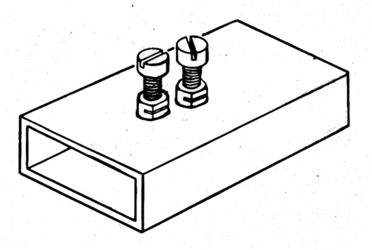
Unfortunately it's not all that easy at times to securely lock the nut after adjustment and, indeed, instability can take place during adjustment.

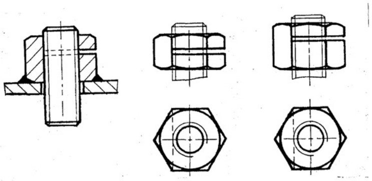
We can use a drop of varnish or to put a complementary nut to secure the assembly. Detuning sometimes occurs after adjustment.

One solution is to use a special nut with a slot in its head. This kind of nut is not commonly available in small nut sizes, so we have to manage ourselves using an ordinary nut, preferably made of brass, and to cut a slot with a junior hacksaw as shown in the diagrams left.

To have the desired effect we have to slightly compress the head in order to get a little tightening when the nut is screwed. In this way we can get good contact during both adjustment and after setting.

In fact, for me, it's easier to do it rather to explain it! Just study the drawings and you'll see what I mean.





Chapter 16

Miscellany

A 10GHz Prescaler

David Wrigley, G6GXK

Background and design aspects
This unit is a practical way of extending the range of a frequency counter. It uses the Fujitsu FMM110VJ chip to divide the input down by a factor of eight. Working over the frequency range of approximately 1 to 12 GHz, it was built to fit directly onto the input socket of an inexpensive frequency counter, successfully extending its range up to 12GHz.

Most low cost frequency measuring equipment currently available won't measure much above 3GHz or so. The Watson F-128 is typical of the low cost units and is good up to 2.8GHz and cost about £60. It works well and is reasonably accurate (it can be calibrated and zeroed in to a frequency standard).

For 10GHz the author used to rely on a cavity wavemeter which, with much eyestrain, is able to measure frequency to within 2MHz or so over the range 8GHz to 12GHz. This is fine for ensuring transmission is within the band and for setting up TV transmitters and other wide band gear but it is not good enough for setting up narrow band beacons, etc, to more precisely known frequencies. With this prescaler ahead of a frequency counter, measurements can be made up to about 12GHz provided that the input is between 0.1mW and 10mW (the chip has an absolute maximum rating of 20mW input power so there is a bit in hand). The accuracy of the overall system is totally dependent on the basic accuracy of the counter used – even a fairly ordinary crystal oscillator is accurate to 10 parts per million. This would give an accuracy at 10GHz to within 100KHz and with careful calibration at a stable temperature the accuracy can be improved to the order of 1KHz. or better - a far cry from the 2MHz accuracy of a cavity wavemeter.

It should be noted here that many frequency counters give a much better display resolution than their accuracy would imply. In other words they display many digits but they all might not all be accurate. This feature can only be checked by calibration against a known frequency standard such as WWV or MSF.

This particular prescaler design was conceived after reading the Dubus Technik IV publication (Ref. 1), which gave details of a design using the Fujitsu FMM110HG package, but at a cost of $300. The author contacted Fujitsu (Ref. 2) who kindly provided information and current prices for this and another smaller cheaper package with similar performance – the FMM110VJ. This latter unit was available in small quantities for about £20 inc. VAT. This was very encouraging and a couple of them were ordered and work started on laying out the PCB.

The circuit (fig.1) follows that shown in the data sheet for test purposes and a decision made at the outset was that no exotic components would be used if they could be avoided. All the other components were in fact SMDs salvaged from old LNBs or other microwave boards. They were all examined for good solder connections and checked for value prior to fitting.

New SMDs can be obtained from Maplin in 25-off minimum quantities or from other sources—they really aren't expensive but keeping costs down is in the authors blood and part of the fun of amateur radio. The circuit is laid out on a tiny PCB using ordinary glass fibre

Figure 1: 10GHz prescaler

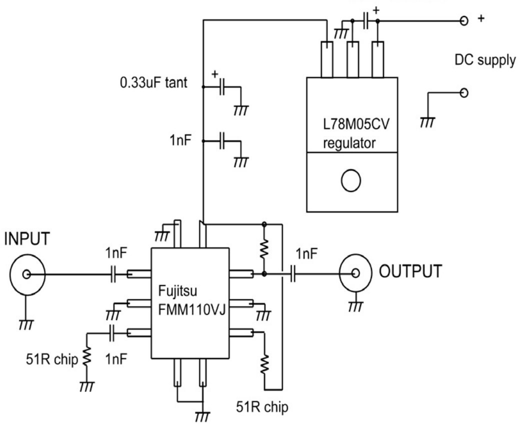


Figure 2: **PCB ARTWORK (ACTUAL BOARD LENGTH = 30mm)**

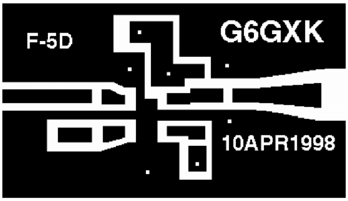
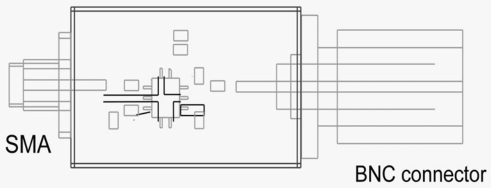


Figure 3: Physical layout



insulated, double sided, copper clad board about 1.5mm thick. The PCB layout is shown in figure 2 (actual length 30mm). The other side of the PCB is used as a ground plane and is almost all copper with only clearances being required for the 5 volt link and connection – these can be cleared with a drill – the ground plane is therefore not etched. It was decided to put in four link pins, using short lengths of wire to bond the earthy parts of the component side to the ground plane as in good microwave practice.

The overall assembly can be seen in figure 3 and is governed mainly by the size of the BNC connector to the frequency counter. The box was formed from 0.5mm tinplate. It was originally thought that the prescaler could have been fed with 5 volts from the frequency counter's internal battery powered regulator, which would have made a very convenient self-contained unit. However it was realised that the chip's requirement of 120-130mA would result in overloading of the internal regulator. A separate low dropout regulator the L4931CV50 was specified but was unobtainable at the time and a LM78M05CV was used instead.

This has worked just fine when supplied from a small PSU at any thing from 7 to 13 volts. It should be noted that the Fujitsu chip gets fairly hot when in use – it is dissipating around 0.65 Watts in that tiny case. For this reason and in order to obtain good grounding it is wise to ensure that it is soldered down to the PCB and that both sides of the PCB are soldered to the tinplate box. The regulator chip will also get hot, especially if it is fed from 10 volts or more, so it is best to mount it on the tinplate box. The prototype had its regulator bolted to a piece of tinplate soldered across the back of the box.

Construction

The PCB is small and simple enough to be cut with a modeller's knife, if an etching facility is not to hand. There should be nothing very critical about the tracks provided the basic layout is followed. Drilling the PCB was accomplished by hand using a 0.8mm PCB drill (ex-radio rally) mounted in the chuck of a pin vice. With a sharp drill this is a speedy process.

The author's eyesight is not as good as it used to be and he has found it necessary to use a headband mounted binocular magnifier to carry out fine work such as this. These are useful devices and are strongly recommended.

Since starting to construct microwave units the author has made himself a low electrostatic field assembly area which consists of a sheet of 0.5mm tinplate over the working surface, with wires soldered to it, to the grounding point of a low voltage temperature controlled soldering iron and to a wrist strap. This was used to assemble the unit and has been used successfully in the past to assemble discrete GaAs FET devices. Of course, modern IC's are tremendously robust but it isn't wise to take unnecessary risks.

Mounting the Fujitsu Integrated Circuit

The first component to be mounted onto the PCB was the IC package. It should be noted that the Fujitsu IC package is very small and the connections are very close to adjacent connections and also to the case, which is ground. It is very easy to bridge these connections with solder, or by poor alignment to the wrong part of the PCB, either of which would of course prevent the chip from working. The important points are:-

- Mount the PCB in a small vice or other holding device to keep it steady.

- Try the IC in place and make sure that the IC connections, which go under the IC package, are clear of other tracks.
- Tin both the underside of the IC package and the PCB area and connections leaving the minimum of solder on the PCB. Use solder wick to remove any excess.
- Use a clean finely pointed temperature controlled soldering iron with the minimum of solder on the tip.
- Hold the IC in place with tweezers, carefully aligning all the pins and gently dab one of the ground pins to attach it to the pcb at its tip. You will find the top left one is the most convenient in the long run. This will locate the IC, minimise misalignment and provide sufficient flexibility to enable the underside of its body to be soldered down.
- Hold the iron close to the chip touching the PCB and the two ground pins which are together at the bottom of the IC so that the heat will melt the solder under the chip.

Press it down with tweezers when the solder is melted and hold for a few seconds until set. It is important that the IC doesn't move out of alignment during this part. The pin connections can now be completed using very fine cored solder wire and with the iron away from the IC to avoid bridging. Use solder wick to clear away excess solder if things go wrong.

After successfully mounting the Fujitsu IC, the rest of the components can be mounted, along with the wire links.

Sourcing the tinplate

Making a tinplate box is fairly straightforward once you have the tinplate. Getting tinplate in Manchester is not so simple. Some tinplate was required to complete another microwave project. After ringing all round the local stockholders with no success, desperation set in and the thickness of old one gallon oil cans was checked – 0.22mm – too thin. So British Steel HQ in London were contacted who, through their Manufacturing Division in South Wales eventually located Lancaster & Winter in Bradford (Ref. 3). Success! So now some of this wonderful stuff was in the author's hands - every radio amateur constructor should have some in stock. Firstly it's much cheaper than brass, it solders beautifully, it can be cut without distortion with my wife's kitchen scissors (but only whilst she's not watching!). It makes good, rigid boxes to any size your heart desires. It not only provides electrostatic shielding - the steel in it also provides magnetic shielding.

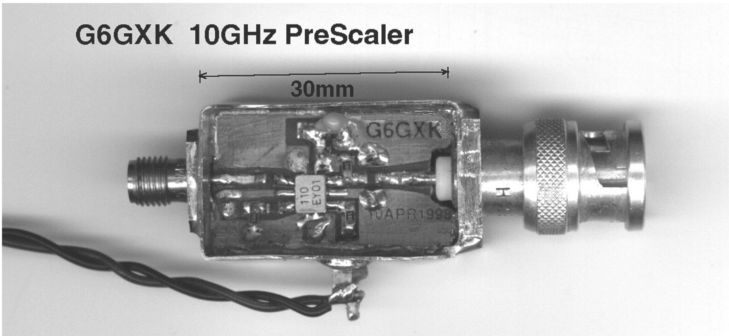
Completing the assembly

The DC supply is fed through a 1nF feed-through capacitor soldered into the tinplate wall. Alongside this the return lead is soldered to a bent up 6BA solder tag, soldered to the tinplate wall. At the back of the PCB the regulator was bolted with cut down leads (4mm) to a piece of tinplate which was soldered across the box so that the leads of the regulator were close to the feed-through can and the 5 volt connection to the PCB. It is important to fit the 0.1uF capacitor across the input to the regulator – otherwise it may oscillate. It was soldered directly across the regulator leads. The output of the regulator goes to the 5 volt input and this is then linked across the back of the PCB to the other 5 volt point.

Testing

The unit is very easy to test. Firstly, spend some time visually checking the soldered connections.

Apply power and check current consumption (about 130 mA). With no input there will be a high level of output at about 870 MHz. This is because the chip oscillates when not driven and this is normal - just remember that if



you think that you are seeing 6GHz or so it may be because the output has fallen to too low a level to trigger the prescaler. Find a source of RF at a level of about 1mW and frequency of 800MHz or more. Apply this to the input and check the counter. It should work without any problem, giving a reading of one eighth of the input frequency. It should also be checked using a 10GHz source such as a Gunn diode. For measurements like this, one useful technique is to use an SMA to waveguide transition and then to a small die cast horn. This can then be pointed at the end of a waveguide or at a horn feed and adjusted until a good steady reading is obtained.

Conclusions

Overall it has been a very satisfying project which continues to prove its worth on a daily basis. As Lord Kelvin is reputed to have said "If you can't measure it then you can know nothing about it!." Modern versions tend to be more forthright. Two prototypes have been built so far with a third one in process. No problems have been experienced in their operation. There are some limitations and possible improvements:

1. It needs to remembered that a strong reading of about 870MHz (6900MHz) will most likely be a "no signal" or "low signal" input condition. As the signal falls off the effect is to move fairly quickly to the 870MHz condition but it doesn't jump there.
2. You need a calculator to multiply the result by 8, not quite as convenient as a direct reading device. There is no easy solution unless the counter you use has provision for prescaler multiples to be input. Any other solution would either require extensive modification to the counter or a worsened resolution.
3. The power input limitations could be improved by using an input buffer amplifier which could both improve sensitivity and provide some limiting against higher powers. Perhaps two opposing diodes across the input would provide some protection against possible overload as available power levels increase.

References:

Ref .1: A 12GHZ Prescaler 1:8.by Angel Vilaseca, HB9SLV and Serge Riviere, F1JSR, DUBUS TECHNIK IV 1995.

Ref. 2: Fujitsu Microelectronics Ltd, Compound Semiconductor Division, Network House, Norreys Drive, Maidenhead, Berkshire, SL6 4FJ. Tel 01628 504800, Fax: 01628 504888.

Ref. 3: Source of 0.5mm tinplate sheets approx 860mm by 800mm about £5.00 each (as at June 98): Lancaster & Winter Ltd , Steel Stockholders, Bradford, West Yorkshire,
BD8 9AE, Tel: 01274 498454

A 10GHz Divide-by-10 Prescaler

Grant Hodgson, G8UBN

Introduction

A prescaler is a digital frequency divider; the output signal is simply the input signal divided by an integer (i.e. a whole number). Prescalers are digital devices and therefore the amplitude of the output is constant and bears no resemblance to the amplitude of the input signal.

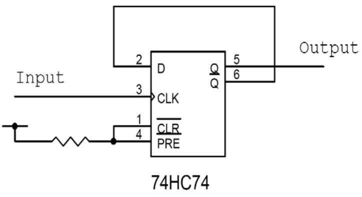
Prescalers have two main functions for amateur microwave use: to extend the range of frequency counters and for dividing the output of a Voltage Controlled Oscillator when used in a frequency synthesiser. This article will concentrate on the former application, although the same design can be used in the latter application as well.

Digital Prescalers

Dividing the frequency of a signal by two is very easy; all that is required is a single D-type flip-flop (or latch) with the input signal being connected to the clock pin and the Q' output being connected to the D input as shown in **figure 1**. The output signal is usually taken from the Q output.

Cascading prescalers (i.e. connecting them in series) is also easy; in this way it is possible to generate division ratios of 4,8,16 etc; division by 64 and 256 is very common. Prescalers operating at microwave frequencies are now commonplace; semiconductor manufacturers such as Fujitsu and Hittite have been making divide by 8 prescalers that can be used with input frequencies exceeding 10GHz for several years [1], [2].

Flip flop used as divide by 2



Low cost frequency counters are now readily available that will operate at frequencies up to 1GHz or even 3GHz, the quality varies somewhat and generally speaking, you get what you pay for. However, microwave counters operating up to 10GHz or above are very expensive, even on the second-hand market, and are often difficult to find at all.

By combining a divide by 8 prescaler with a frequency counter operating up to (for example) 1.5GHz, it is possible to measure the frequency of a 10GHz signal, and possibly being usable to 12GHz. However, in order to determine the exact frequency of the device under test it is necessary to multiply the reading on the frequency counter by 8 - not too difficult if the counter reads 1.1101101, but multiplying a displayed number such as 1.2778563 is a little more complicated, usually requiring a calculator.

New Divide by 10 prescaler

Thanks to some recent advances in high-speed digital electronics, the problem of multiplying the displayed frequency on the counter by a factor of 8 has been solved. Hittite Microwave Corporation [3] have recently (October 2002) released the HMC438 which is a revolutionary divide by 5 prescaler IC. This remarkable little IC requires only a single 5V supply and a couple of external capacitors to operate; the input frequency extends from DC to 7GHz. The internal circuitry of a divide by 5 prescaler is a trivial task when working at low frequencies using standard CMOS techniques, but when working at microwave frequencies the number of individual transistors required to form the appropriate circuitry poses a number of technical problems, and the availability of the HMC438 marks a significant breakthrough.

Hittite also make a divide by 2 prescaler which operates from DC to 11GHz; cascading these devices results in a true divide by 10 prescaler that can be used up to at least 10GHz and beyond. The full circuit diagram is shown in **figure 2.**

All three ICs are powered from a single +5V supply, and a 5V regulator is fitted to the PCB but not shown on the schematic for clarity. This allows the prescaler to be used with a standard 12 – 13.8V DC supply, and gives some protection against accidental polarity reversal.

The input signal is DC blocked by the capacitor C1. The value of this capacitor determines the sensitivity at both high and low input frequencies. Due to the fact that the prescaler is intended for high frequency operation, the value of the capacitor has been chosen to maximise the input sensitivity at higher frequencies.

The input signal is amplified by IC1. This is a Gali-1 MMIC (Monolithic Microwave Integrated Circuit) from Mini-Circuits. For input signals greater than approximately –10dBm this amplifier is driven into saturation, which ensures that the output level is constant. IC2 also provides a secondary function in the form of an input protection buffer; a large input signal will simply be limited by IC1, thus giving a constant level output signal. If an excessively large input signal is applied IC1 may be damaged, but this is much easier and cheaper to replace than IC2.

The level of the signal is then attenuated by the Pi-attenuator R1-R3. The signal level at the output of the attenuator is at a level of approximately -2dBm which is close to the optimum level for IC2. IC2 is the first prescaler which is an HMC361S8G. This divides the signal by a factor of 2.

The HMC361S8G has two balanced inputs; the attenuated signal from IC1 is connected to one input (pin 5) and the other input is connected to ground via the capacitor C12. The HMC361S8G also has two complementary outputs; again only one of which is used – the other is simply left open circuit. The divide-by 2 output at pin 3 is DC blocked and connected to one of the inputs of IC3, an HMC438 divide by 5 prescaler. As with IC2, there is an unused input which is grounded by a capacitor and an unused output which is left open circuit. The output of IC3 is at pin 7 which is DC blocked by C11; this output signal is exactly 1/10th of the input frequency and can be connected to any suitable frequency counter. The level of the output signal is approximately –1dBm.

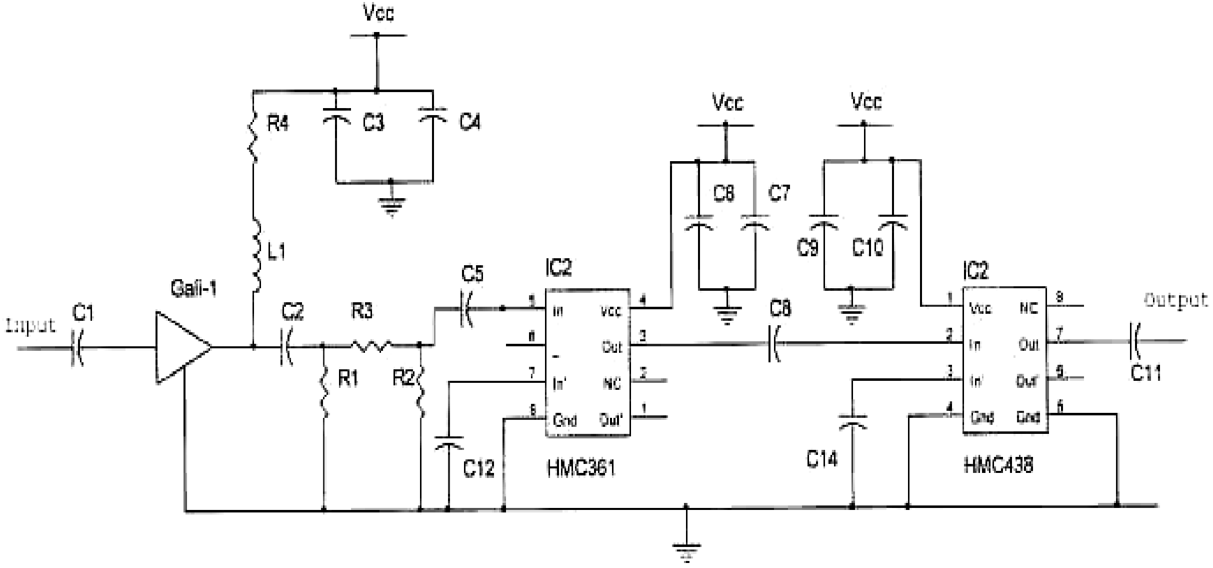
IC1-3 are connected to a single +5V supply; both prescaler ICs have two decoupling capacitors placed close to the VCC supply pins. L1 is a Mini-Circuits ADCH-80A broadband choke which provides a high inductive reactance from 50MHz to 10GHz – quite a remarkable performance in it’s own right! R4 sets the current through IC1 to 25mA.

Construction

A PCB has been made for this project for those that feel confident to be able to solder the small devices. The PCB has provision for all the components including a 5V regulator and SMA ‘end launch’ sockets for the input and output. The PCB also has provision for a number of other features that will be briefly described later.

The biggest problem when trying to

Figure 2 – circuit diagram



build this particular project is soldering the prescaler ICs, and in particular the HMC438 which is very small indeed. Both the HMC361S8G and the HMC438 have a ground 'slug' on the underside of the package that cannot be seen when the IC has been soldered. The intention of the IC manufacturers is that these components are soldered using commercial SMD reflow techniques whereby solder is applied automatically in paste form, and melted in a special oven. These techniques are not suitable for small production runs, and the equipment required costs about as much as a small house. However, it is possible for advanced constructors to solder these devices at home, although some experience with small surface mounted components is required, as is some form of optical aid, a heat gun and some solder paste.

For best results the prescaler ICs are soldered first, one at a time. The technique involves placing a small amount of solder paste on the central ground pad on the PCB, and either applying a small amount of solder on each of the PCB pads for the IC pins, or applying the solder paste in a long line for pins 1-4 and 5-8. The prescaler IC is then very carefully placed onto the PCB, taking care to note the correct orientation. The IC will sit on top of the solder paste, and it helps if it is pushed down very slightly without twisting or moving it. The solder paste is then melted using a heat gun; the most suitable tool is a small, high wattage (> 1000W) gun used for heatshrink tubing with a small nozzle. Extreme care must be taken to ensure that the correct amount of heat is applied – too little heat and the solder paste will not have melted, which can lead to small solder balls which in turn can cause short circuits. Too much heat will damage the PCB and/or the IC. The right amount of heat will melt the solder paste properly, simultaneously soldering all 8 pins and the ground slug. As the heat is applied, several things happen :-

1) The solvents in the solder paste evaporate and the flux becomes active.
2) The solder on the pads for the IC pins melts. As this happens, the surface tension of the liquid solder pulls the IC on each of the 8 pins. If the solder on all 8 pins melts at approximately the same time, the IC will automatically be pulled to the exact centre of the pads – even if it was placed with a slight offset. The effect of this has to be seen to be believed – it really does look like magic, but is really just the application of physics!
3) As the solder melts, it naturally flows onto the exposed, tinned pads on the PCB, and so any paste that has been applied onto the areas covered by solder resist (the green coating on the PCB) will tend to flow towards the nearest exposed pad, thus automatically reducing the chances of a short circuit between adjacent pins.
4) At this point, the solder paste on the underside of the ground slug has not fully melted. It is necessary to keep the heat applied to the IC whilst the solder on the outer pins is still molten.
5) Then the IC will move slightly downwards as the solder under the IC melts and the surface tension pulls the IC further down onto the PCB. This is a very subtle effect, but can be seen with some experience and especially with good optical aid such as a microscope.
6) At this point the heat is removed and the board is left to cool, and then the solder paste for the other prescaler can be applied and soldered as above.

There has been some considerable debate on the US Microwave reflector [4] recently about alternative methods of soldering devices such as the Hittite prescalers with the 'hidden' ground slug; alternatives to the use of a heat gun are to place the PCB on a hotplate at a temperature considerably greater than the melting point of solder, or to use a conductive epoxy to mount the prescaler ICs. The use of epoxy would be an option, but unlike solder paste is very difficult to obtain in very small (i.e. cost-effective) quantities, and requires special care when curing, so it is felt that the use of solder paste would be the best option for home construction.

Note that the HMC438 is considerably smaller than the HMC361S8G – the HMC438 has pins on a pitch of only 0.65mm – that is the distance between the centre of the pins, not the gap between them! Although very small, this is an industry standard package, and is widely used for many ICs – especially microwave ICs. The use of good optical aid is mandatory; there has been discussion of this recently both on the US microwave reflector and in Technical Topics in Radcom.

Some of the RF coupling and decoupling devices are also very small, being of 0603 size – i.e. 1.5mm long x 0.75mm wide. These components can be soldered with a soldering iron with a small tip, fine solder (preferably 30SWG, although 26SWG can be used), optical aid and of course a steady hand! The reason for using such small capacitors is that they have a much lower self inductance, and therefore a higher self-resonant frequency. This means that higher values of capacitor can be used (4p7) for the high frequency part of the circuit, which increases the sensitivity at lower frequencies such as 4GHz. Therefore, smaller capacitors have the effect of increasing the effective frequency range over which the prescaler can be operated. It is fully appreciated that some microwave constructors will feel somewhat nervous about performing such delicate soldering techniques on tiny, expensive devices. For this reason the author is considering the option of making the PCB available with the ICs already soldered and tested, and possibly some of the other small surface mounted devices mounted as well. This would give a 'half-way house' whereby the hardest part of the construction has already been done, but the easier parts such as soldering the connectors and IC1, and mounting in a case would still be done by the constructor.

Printed Circuit Board

The PCB for this project requires special mention. It is made from 1.6mm FR4 (fibreglass), but instead of having the components on one side and a ground plane on the other, the PCB has 4 layers, with 2 ground planes sandwiched in the middle of the board. The reverse side of the PCB is thus free for extra circuitry, and has been used as a 'Microwave Experimenter's Project Board' which consists of pads for two Mini-Circuits Gali-MMICs, a passive, broadband frequency doubler, a SPST RF solid state switch and the possibility of a second frequency doubler. The use of a 4-layer PCB is believed to be unique in amateur microwaves. It has the advantage that although the RF properties of FR4 are inferior to Duroid and equivalents, the ground planes are only 0.3mm below the surface layers, which may allow the use of this type of board at 24GHz.

Of course the losses in the FR4 will preclude the use of this type of PCB for LNAs and PAs at such frequencies, but for other purposes such as doublers, mixers, filters and driver/buffer amplifiers this new 4-layer PCB technology may be usable at a much lower cost than Duroid, and can be manufactured commercially with plated through holes (vias) in very small quantities.

Any losses can easily be overcome with new high-frequency MMICs (such as the Galis), (at 10GHz or 12GHz) and 1.6mm

FR4 is considerably stronger than 0.25mm (or similar) PTFE material.

Performance

The maximum input frequency at which the prescaler will work is determined by a number of factors, including the gain of IC1, the loss associated with L1, the reactance of C1, C2 and C5 at frequencies above 10GHz and the upper frequency limit of IC1. In practice, the prescaler has been found to have an upper frequency limit of around 14GHz at an input level of +13dBm. IC2 has a specified maximum frequency of 11GHz, so this extra performance should be considered a bonus and cannot be guaranteed. For use at 10.4GHz, the minimum input level is approximately –15dBm, which makes the prescaler very sensitive. The maximum input level is 15dBm which is the absolute maximum input power for the Gali-1 MMIC. The prescaler has a definite cut off point; when the input signal is even very slightly above the maximum operating frequency the prescaler simply stops working and the output becomes unstable. Therefore it is very evident whether or not the prescaler is working properly. Note that this is usually not the case with a frequency

counter, where the counter usually starts to display a frequency slightly less than the true input signal. Also note that when no signal is present a the prescaler input, the prescaler becomes highly unstable and oscillates. However, unlike the Fujitsu divide by 8 prescaler (which oscillates at a fairly constant frequency with no input signal), the Hittite prescalers seem to produce a number of spectral lines which cannot be resolved by some frequency counters, although the may be harmonics of the fundamental oscillating frequency. This instability is to be expected, and is believed to be caused by the positive feedback action of the input circuitry which tries to 'capture' the input (sine wave) signal. With no input signal present, the input circuitry of the prescaler is trying to capture noise and becomes unstable. However, as soon as an input signal of sufficient level is present, all signs of instability disappear and the prescaler behaves normally. The DC current drawn by the prescaler does not increase in the unstable (no input signal) state.

Options

It is fully appreciated that this is not a cheap project, although it is felt to give reasonable value for money considering the high performance and the use of the newest prescaler technology. For those that want a 10GHz prescaler at a lower cost, IC2 can be replaced with an HMC363S8G divide by 8 prescaler IC, IC3 and associated components are not fitted and the output from IC2 routed directly to the output socket.

For an even higher frequency prescaler, IC2 can be replaced with an HMC364S8G which is a divide by 2 prescaler identical to the HMC361S8G, but with an upper frequency limit of at least 13GHz, and may be useable well beyond that, although this has not yet been tested.

For more sensitivity at 10 or 12GHz, IC1 could be replaced with a Gali-19 or Gali-19 high frequency MMIC; the author is currently testing this option.

Parts List

C1,2,5,12	4p7
C3, 11	100pF
C4, 6, 9	10nF
C7, 10	10uF Tant
C8,14	10pF
R1, 2	220R
R3	22R
IC1	Gali-1
IC2	HMC361S8G
IC3	HMC438

References

[1] 12GHz Prescaler, David Wrigley G6GXK, CQ-TV 185 (published by the British Amateur Television Club)
[2] New 12GHz Prescaler using Hittite HMC363S8G, David Wrigley G6GXK, www.microwave.fsnet.co.uk/projects/projects-1.htm
[3] www.hittite.com
[4] www.wa1mba.org/reflect.htm

Circular Waveguide - the Truth!

Richard T. Nadle, K2RIW

Editor's comment: *This article was written in 1999 after a series of practical experiments undertaken by Dick and his friends on Long Island. It appeared in print in the North Texas Microwave Society newsletter "Feed Point" in August 1999 and was also disseminated among those subscribing to WA1MBA's internet microwave reflector. Our thanks go to Dick for giving us permission to reprint this most interesting information. The 3/4" pipe Dick used in these experiments is available over here as 22mm (19mm internal diameter) copper pipe.*

1. INTRODUCTION -- I've received 8 e-mails, with some nice words of encouragement, about circular waveguide. There seems to be a thirst for knowledge on this subject -- the "poor man's" high performance WG. Some people wanted the extra decimal places and some wanted more description of the 10GHz 3/4" copper pipe experiments of "The Ten-X Group" on Long Island. In 1997 we burned a lot of "midnight oil" at the QTHs of N2LIV (Bruce, the president) and N2NKJ (Ron) while performing the pipe WG experiments. On the fourth night in three weeks, we again ended our experiments (with blood shot eyes!) as the sun was rising; we then knew we were pretty serious, or a little crazy, and that circular waveguide was great stuff. It was the Shepherd's Crook Dish Feed assembly of the San Bernardino Microwave Society (WA6EXV design, I believe) that got us started in the experiments.

At first, there were a few East Coast microwavers who thought the 3/4" copper pipe WG feed assembly, with four elbows, would have high loss …. we found that it actually has about as low a loss as you can get.

2. CIRCULAR WG FREQUENCIES -- If a 3/4" water pipe had exactly 0.7500 inch inside diameter, it would support the TE11 circular mode (the dominant mode) between the absolute min/max frequencies of 9.225 GHz and 12.045 GHz.

(The UK's 19 mm ID pipe will have cut off frequencies of 9.25 and 12.08 GHz …. editor)

Here are the first 6 cut-off, "absolute" frequencies (no guard bands) for 0.7500" ID pipe:

Frequency (GHz)	**Mode(s), Circular**
9.225 to 12.045	TE11
12.045 to 15.301	TM01
15.301 to 19.192	TE21
19.192 to 21.053	TE01 & TM11
21.053 to 28.002	TE31
Above 28.002	TM21

Some day I'll explain where all the extra digits are coming from. We've devised a way of getting Bessel functions to 12 decimal places; I was on a project that needed that.

3. FREQUENCY SCALING -- Once you know the ID of your "pipe", you can scale all these frequencies. So-called 3/4" water pipe comes in K, L, and M styles where each has a different wall thickness and slightly different ID. There are also multiple kinds of soft copper pipe, as well as some corrugated pipes, that will make great "semi-rigid" waveguide that can be bent around corners. You can measure the exact ID of your pipe with callipers and inversely scale all the above frequencies accordingly (bigger diameter is a lower frequency for each mode). If you intentionally deform the soft copper pipe with a good roller mechanism, you will make "poor man's Ellipto-flex".

Ellipto-flex has a very slight loss increase (for the same circumference) but it forces a particular polarization, and has a slightly wider frequency range for the dominant mode …. More on ellipto-flex at another time.

The mathematics called Bessel functions predict FM modulation and circular WG characteristics. Most of the books on circular WG are not very clear. One kind of Bessel function, of multiple sets, predicts the TM circular modes and the other kind of derivative Bessel, of multiple sets, predicts the TE circular modes. There's got to be a better way that "they" (or we) can explain all this.

4A. BOOK #1 -- The best reference I've found on the subject (I had to read the circular WG section ~ 10 times) is Theodore Moreno's "Microwave Transmission Design Data", Dover Publications, 1948 (original), 1958 reprint, available for $8.00 at ABEbooks.com.

4B. BOOK #2 -- Concerning WG components, here's THE BOOK: George Southworth, "Principles and Applications of Waveguide Transmission", D. Van Nostrand Co., 1950; 689 pages (an oldie but goody). It contains some of the best PICTURES of how rectangular and circular WG really works with lots of performance curves (you won't need the math to understand the pictures (pages 166 & 169), it's almost an animation) -- amazing stuff for 1950. Page 121 (A & B) has pictures are 21 of the circular WG modes (with the relative sizes of pipe shown, same frequency) made with an "RF absorbing camera". The book shows some great transition devices, hybrids, mode killing devices & devices for launching higher modes (pages 354 to 362), round WG components (pages 269, 327 & 328), circular guide fin line (page 133), a great section explaining choke flanges (page 201), a circular pipe polarization rotating device that's "home brewable" (page 207), the shapes of circular and rectangular WG (of constant periphery) that give minimum loss (page 193)(the popular ones are not optimum), "skeleton WG" (page 175), about 15 kinds of WG irises (page 246 & 255), circular WG filters (page 307), the Qualcomm duplexing filter explained (page 309); linear, binomial, Gaussian, and exponential WG impedance stepping functions for broadband impedance matching (pages 269 to 276), 14 designs for dummy loads (pages 368 to 371), about 25 attenuator designs, rotary vane phase shifter (page 333), rotary vane attenuators defined (page 375), a way of designing a variable conductance dissipative film (page 377), 33 pages of horn data (only portions have appeared in other WG or antenna books), 8 kinds of "backfire" feeds including the Cutler (pages 448 to 454), eight types of WG slot antennas (pages 425 & 430), five kinds of corner reflectors, waveguide lens antennas, some TWT and magnetron info, etc. The picture on page 186 shows me how I could make S-band waveguide out of rain gutter down spout tubing. Let a Microwaver stand in a good hardware store with that book in hand and I think he'll get some great and crafty microwave ideas!

5. SAME WAVEGUIDE MODES -- It is interesting to note that the main pipe WG mode, TE11 circular, and the main rectangular WG mode, TE10 rectangular, are the exact same mode. They use a different definition of the subscripts for circular and rectangular that makes them look different but they're not. Once we realize this it becomes easy to construct (HOME BREW!) devices that transition from rectangular to circular. We of the Long Island "Ten-X Group" of microwavers did a fair number of experiments in this area a few ago. Here are some of the results.

6. RECTANGULAR TO CIRCULAR TRANSITIONS -- Our "Elegant Transition" was a carefully constructed slow transition (over a 1 foot length) from rectangular to circular. It had an S11 of -35 dB (VSWR = 1.04). A "sloppy transition" was made by crushing one end of a 3/4" copper pipe in a vice and forcing it into a WR-90 WG flange *(WG16 flange in the UK ..editor)*. It had an S11 of -23 dB (VSWR = 1.15). To make these measurements we used a Super WR-90 (UK WG16) 20 dB coupler (50 dB of directivity)

7. DUMMY LOADS -- In circular WG, these are quite easy to construct. Simply sharpen a 3/4" broom stick handle and force it into the 3/4" copper pipe. About 3" of taper and 2" of non-taper is fine. The usual moisture in the wood makes a great "slow absorber", which makes it more forgiving of errors. The main difference between a -35 dB S11 dummy load (VSWR = 1.04, [sharp tip]) and a -20 dB S11 (VSWR = 1.22) seems to be how sharp the point was at the tip of the broom stick handle and was the taper too abrupt (too short). There may be some variations caused by knots in the wood, but we didn't seem to have that problem.

The completed circular WG dummy load consists of a ~ 7" piece of 3/4" pipe with the tapered broom stick handle (absorber) in it plus a copper pipe coupler at the open end. Some of the broom stick absorber can stick out the pipe far end, if you prefer. It is easy to place this load on any other piece of circular WG, while running component tests. These pipe couplers really are "sexless" connectors. For experienced rectangular WG users, it will feel strange to make connections in 2 seconds and not worry about screwing down the flanges to get a good VSWR!

8. LAUNCHERS -- A circular WG to SMA launcher consisted of a 3/4" copper pipe end cap, soldered onto ~ 1.5" length of 3/4" copper pipe and an SMA Female to Female (bullet) is fed through a hole in the side of the end cap plus pipe, ~ 1/4 wave away from the closed end (it is either threaded into the end cap and use a nut, or simply solder it). A ~ 1/4 wave long wire probe inside does the launching. A pipe coupler is slipped over the 1.5" length of pipe that protrudes. The completed launcher can be placed on any piece of circular WG in 2 seconds by the "sexless" connector (pipe coupling) method.

Proper adjustment of the position, tilt (forward and aft), and length of the probe wire is a bit tedious. Place a circular WG load on the launcher output and measure the VSWR at the SMA connector to find the proper probe length; it's an unambiguous operation. Once you get the hang of it you will make quite a number of launchers at once. The one component of the launcher that costs anything is the SMA bullet.

9. PADS -- We never did this but it would be easy to design circular WG fixed attenuators by decreasing the length of broom stick absorber and tapering both ends to have a good impedance match from either direction. In this case I would recommend painting the absorber to keep the moisture content (absorption) constant. If it is found that the loss is too great for a convenient length of tapered wood absorber, consider making the absorber out of six "splines" by using thin sheets of wood, or out of balsa wood. These low density materials (with tapered ends) will allow a lower insertion loss to be constructed from a longer length of wood absorber. Also, the slower loss characteristic will cause a lower VSWR for a particular taper rate.

10. COUPLER (SEXLESS CONNECTOR) -- There is a kind of copper pipe coupler that has no internal ridges or dimples. These will allow the 3/4" pipes to directly touch(butt against each other inside the coupler) without any gap. A gap will have a larger inside diameter, within the coupler. Even with a gap, the VSWR impact was very small. If these couplers are unavailable, a purist can remove the ridges or dimples with a round file or an adjustable reaming tool. However, you will find that the moulded copper fittings are hardened and the reaming/filing operation is a quite difficult.

11. OTHER COMPONENTS -- We discovered that many of the hardware store 3/4" plumbing fittings were almost "designed" for us microwavers. A 45 degree elbow and a 90 degree elbow gave us VSWRs of ~ 1.3 and 1.4, as I remember. The copper pipe "coupler" makes a great trombone and polarization twister. The end caps are a great way of constructing 3/4" circular WG to SMA launchers. When the circular WG components are properly plugged together with these "sexless" connectors, we haven't found any fittings that change significantly in VSWR as you do or do not solder them.

Test everything by plugging them together, solder them later. Soldering seems to only be required for mechanical reasons.

You will find that many circular WG components are **VSWR sensitive to rotation** (at the sexless connectors). Many circular WG objects can create elliptical polarization. Southworth's book covers this subject on page 206. A liberal sprinkling of septums within launchers, horns, etc. will force a linear polarization at the output point (usually without loss) and they insure that the elbows and other non-symmetric components don't create elliptical polarization. A septum consists of a thin sheet of metal soldered across the diameter of the circular WG at right angles to the desired polarization.

A screw protruding into the guide makes a great way of correcting VSWR; however, depending on its polarization, it could also create some elliptical polarization -- so use a septum in the vicinity of the circular WG output. Find the best location for the screw by using a steel BB inside the WG and positioning it with a magnet on the outside.

Copper transitions from 3/4" to 1" and 3/4" to 1.5" make nice feed horns. A 3/4" soldered-pipe to threaded-pipe adapter makes a convenient flange for the centre of a dish antenna. That's where you feed the Shepherd's Crook 3/4" circular WG through the dish. Two large washers and a large nut can hold this assembly in the dish centre. It is also possible to mount a copper 4 hole flange to the back of the dish centre and feed the round WG through it. In each of these cases it will be necessary to ream out the shoulder within the fitting to allow the 3/4" pipe to pass through the fitting. We purchased a mechanically adjustable ream to accomplish this. The operation was quite difficult because of the hardening of the moulded fittings. With a lathe this operation was a lot easier, when there was enough metal to mount in the lathe chuck.

12. WG POLARIZATION -- I must give you a warning! If you are using a long length of pipe as WG up your tower (K2TXB and W2DRZ did this), the ellipticity of the pipe (the "run out") could cause a horizontal polarization, launched at the bottom, to become a vertical polarization at the top of the tower (the polarization can rotate). By simply rotating the launchers at the top and bottom of the WG, you will find the lowest loss combination for that "pipe WG"
There is also this added danger ... when you are in your Home Depot Hardware Store, with your callipers, while buying 10' lengths of pipe that

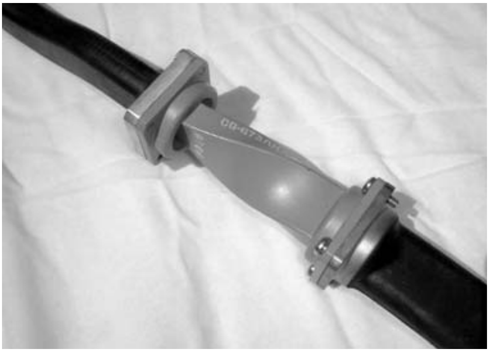
have no run out, you will find the following ... one pipe in 3 has almost no run out, that's the ones you want to buy; (2) The Plumbing Department Foreman will watch you during this process and he will say, "Oh no, another one of those crazy engineers who's doing a precision plumbing job in his house!" You will have to suffer some ridicule in the pursuit of your art. You probably will not succeed in explaining what you are really doing, just remember -- you're tough and can take this ridicule, you are pursuing a far off goal, that great 3cm DX!

13. CIRCULAR WG DISH FEEDS -- For a 0.6 F/D dish, the W2IMU Dick Turrin dual mode horn intentionally launches the TE11 and TM11 circular modes at just the right amplitude ratio and relative phase. That's what gives it such a beautiful pattern ... No edge currents, no side lobes and great control of the electronic phase centre versus azimuth, elevation, or diagonal scans. Any shift in the electronic phase centre of a horn versus observation angle creates the same antenna pattern degradation as if you had the same amount of error (like a big dent) in that area of the parabolic reflector it is illuminating. With this understanding, it becomes more obvious why this horn achieves higher dish efficiency. For deeper dishes (~ 0.3 F/D) the Scalar feed (the one with the multiple rings) does almost the same job.
But, be careful. The W2IMU multimode circular horn can get a Microwaver involved in an altercation or a law suit. I constructed my 2287.5 MHz W2IMU horn from food cans for Apollo 15 and 16 reception "Houston This is Apollo" (QST June 1972) and "A 12 Foot Stressed Parabolic Dish Antenna" (QST August 1972). I measured the diameter of all the food cans in the local super market. The store manager became alarmed. and asked me to leave. He thought I was a Ralph Nader advocate and was about to start a litigation (I had to endure ridicule way back then too). I settled on Scotts Oatmeal cans, purchased from a Foods of All Nations Store; they had the correct diameter for the launcher and RHC polarization section.

14. CONCLUSION -- I hope you all enjoy the above info. There are a few more 10.368 GHz 3/4" copper pipe experiments to be described later. Please feel free to correct the errors or add to the info.

Making more Flexible 90 degree twists in Waveguide

Kent Britain, WA5VJB

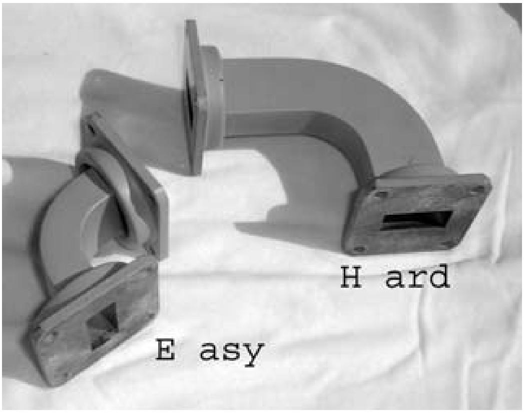


Getting waveguide to go around a rotator, or even to be flexible going up a portable mast, can be fun.

Waveguide has E and H planes. One way to remember which is which, is it to think of these planes as H hard and Easy to bend.

You could easily bend waveguide along the Easy Plane, but it's a lot Harder to bend it along the H bend. It's that Easy and Hard planes that make it fun to get even the 'Flex' guides to twist. They bend easy one way, but not the other.

When it come to E and H bends in waveguide, just look at it and think which would be easier to bend, that's the E bend. Keep an eye out for 90 deg waveguide twists. Using two short sections of flexiguide with a 90 deg twist in the middle makes the whole section much more flexible.



Got the Connection?

Richard T. Nadle, K2RIW

Concerning microwave connectors, the story goes ….

N is for Neil**, C** is for Councilman. Those are the connector names.

BNC is Bayonet Neil Councilman, **TNC** is Threaded Neil Councilman.

Likewise, I was told **SMA** was Sub Miniature type "A". **SMB** was Sub Miniature type "B" and **SMC** was Sub Miniature type "C", all originally Bendix names (they invented the SMA).

I did not originate this story; I got it from an older issue of the "Cheese Bits", the monthly journal of the Mount Airy VHF Radio Club (The Pack Rats), from the Philadelphia, PA area. I've repeated the story to many RF people and have encountered older engineers who remember the creation of those connectors. Sometimes they confirm the story, so it might be true.

Most people who served in the US Navy have been told that the N Connector stands for Navy Connector, instead of the Paul Neill Connector. I've worked in the RF and Microwave world for over 35 years. RF people use coaxial connectors every day and almost every other day they hear the word RADAR. To me it's amazing that most engineers never ask what these abbreviations stand for. When you tell them, they are amazed that the letters really stand for something, as apposed to being randomly chosen letters such as C Band, S Band, or V Band.

Here is a way you can have some fun with your fellow engineers. Just ask them what do the letters stand for in N, C, BNC, TNC, SMA, SC, HN, RADAR, or SONAR. You will probably find that less than one engineer in 10 knows what ANY of those abbreviations stand for, regardless of their years of experience.

I've run that experiment at AIL, Kmec, Rockwell, Northrup, Sanders, Cutler Hammer, MPD, Narda, Telephonics, Bell Labs, Johns Hopkins APL, Edwards AFB, Wright Paterson AFB, Eglin AFB, NASA Goldstone, NASA Goddard, NASA Kennedy, NASA Johnson, China Lake NAWC, Pt. Mugu NAWC, Johnstown NAWC, Rome ADC, Bell Northern Research (Canada), Westinghouse, Comtech, Bendix, Lockheed Martin, AEL, IBM, Microwave Associates, Amplica, HP, Agilent, Avantek, Mitec, Eaton, EDO, and a few more! The ratio is always the same and the amazement of the engineers is always the same. The information seems to be a well kept secret and I wonder how it happened.

I'm saddened that the engineers who created these great connectors are such unsung heroes. There are companies (such as AIL) where almost every product, for the last 55 years, has something to do with RADAR -- Receivers, Transmitters, Exciters, Jammers, Antennas, Direction Finders, Surveillance, Reconnaissance, Diplexers, DSP, Range Doppler Matrix, Pulse Forming Networks, Rotary Joints, Pulse Compression Networks, Synthetic Aperture RADAR (SAR), Inverse SAR, TWTs, HPAs, LNAs, etc. Choose an engineer who has worked there for 20 years (or any other number) and tell him the following: "RADAR has put food on your table for the last 20 years, it's one of the most important words in your vocabulary; what does it stand for?" You will see him struggle to find a word for the first A in RADAR

-- there isn't one, it's not a real acronym. RADAR is a contraction that stands for RAdio Detection And Ranging -- according to Skolnik, "Introduction to Radar Systems", 1962, page 1. While he is struggling to come up with the definition I will say, "you know, Pete in the Transmitter Department knows RADAR so well that he can even spell it backwards". I once asked that question of 15 engineers at a RADAR savvy company before getting the correct answer. It was the 5th engineer in the RADAR Department who knew the answer!

Before you try any of these tricks on your fellow workers, be sure you have chosen someone who likes you and has a reasonable sense of humour. Some people take themselves so seriously that they can't stand to have any of their weaknesses exposed without blaming you, the exposer.

Have fun with this information but be careful -- there is a mine field out there of people who can't take a joke concerning their profession. We all make mistakes, and we all have missing pieces of information -- try to enjoy the educational process. If you can't laugh at yourself, you're

probably taking things too seriously and living a much harder life than you have to. Mother Nature didn't make this RF Universe in such a way that it's easy for anyone to really understand a whole lot of it. It's probably that challenge that got most us into the Microwave business/hobby in the first place.

By the way, that beautiful little (~ 2mm) push-on RF connector that's located on your Wireless LAN Card (it's a closed-circuit jack intended for an external antenna) is an **MMCX** Connector (Micro-Mate C). There are different companies that sell a gold plated SMA to MMCX (Male) adapter

for prices that range from about $8.00 to about $60.00 each. The quality seems to be the same. There are experimenters that have gotten miles of 802.11b DX with an external 2.4 GHz Dish antenna connected to their barefoot Wireless LAN Card.

RF Cables and Connectors Don’t Get Any Respect … But They Should!

From an abridged article by Jim Pomager, Editor-in-Chief RF Globalnet

***Editor’s note:** Our thanks go to **Jim Pomager, Editor in Chief at RF Globalnet** for giving permission to use his article.*

Unlike active devices such as ICs, cables and connectors are quite possibly the most unglamorous components in the RF/microwave design world. Like doorknobs in an architectural plan, RF cables and interconnects are rarely considered during the design process itself. Rather, these necessary evils are typically contemplated only after a project is all but complete.

They truly don't get much respect. However, these oft-neglected components can have a significant impact on overall system performance. In the commercial field, RF and microwave design engineers face enormous pressure to deliver quality products at lower costs. As a result, cable and connector vendors can end up being their best friends, helping them meet a variety of electrical, mechanical, and environmental demands. In fact, today's cable and connector suppliers are supporting designers with products that are smaller, faster, and more phase stable than ever before. Microminiaturized products handle higher frequencies, conserve space

There are two key factors behind the continuing reduction in RF cable and connector sizes. First, the frequency ranges of RF/microwave systems have been creeping steadily higher
(driven principally by the need for higher data rates in communication systems) and small diameter coaxial cables are generally more efficient at carrying the higher microwave frequencies than are large-diameter cables. Hence, we have witnessed the rise of small-diameter, high frequency cable -- and the microminiature connector. One line of connectors that illustrates this trend toward higher frequency and smaller size is the MMPX series from Huber+Suhner. These board-mount, microminiature, snap-on connectors are comparable in size to the diminutive MMCX connector and are compatible with the MMCX interface. They demonstrate linear VSWR at frequencies from DC to 65 GHz and have shielding effectiveness of -85 dB from DC to 26.5 GHz, -65 dB from 26.5 to 50 GHz, and -60 dB from 50 to 65 GHz. Applications include test and measurement, defence, and mobile radio. MMPX connectors are available in straight cable plug, straight cable jack, right angle cable plug, straight PCB jack, edge-mount, PCB jack, and adaptor configurations.

For more information, visit
www.hubersuhner.com.
Tensolite brings its SMP and SSMP lines of blind-mate push-on connectors to the microminiature party. The SMP series covers DC to 40 GHz and

permits connector spacing as close as .170 inches for an assortment of military, space, and telecommunications applications.

The SSMP series ratchets it up a notch in frequency (to 65 GHz) and down a notch in size (spacing to .130 inches) for smaller, lighter systems. SSMPs are offered in popular interconnection versions for cable assemblies, surface launch, edge launch, hermetics, adapters, and loads. For more information, you can download Tensolite's RF/Microwave Products Catalogue from their website at

www.tensolite.com

The second major factor influencing RF cable and connector size is the shrinking of the RF/ microwave system itself. With almost all package sizes getting smaller. As a result, cables and interconnects (all components, for that matter) must have minimal footprints, whether they are on a board, in a base station, or on a tower.

Nowhere is this lack of space more evident than inside a wireless device. For this cramped environment, RF Industries introduced the MHF series of micro connectors and cable assemblies for Wi-Fi. These coaxial connectors have a mating height of 2.5 mm. They support dual-band applications in the 3 to 6 GHz range, including laptops, PDAs, and cell phones. Cable assemblies are available with a range of connectors, including SMA, MMCX,TNC, and Type N. For more information, visit

www.rfindustries.com

In the test and measurement arena, where cables are constantly being attached/detached from instrumentation, recent advances in interconnect technology can help save engineers a lot of time (and a little sanity). Winchester Electronics (USA) just released its QC-SMA series of SMA connectors, which cover the frequency range of DC to 6 GHz. These connectors incorporate a push/ pull style of mating, which allows them to be connected and disconnected without the need for special tooling. (This is in direct contrast to standard SMAs, which must be torqued for proper installation). Once QC-SMA connectors are mated, they can rotate 360 degrees without losing connection.

In addition to test environments, these connectors can also be used in microwave subsystems, base stations, mobile radios, and other applications. For more information, download the QC-SMA datasheet at:

www.winchesterelectronics.com/products/qc/qcsma.asp

Stainless steel connectors …. good or what?

Peter Blair, G3LTF

This may only be of interest to microwave EME people! Some days back I asked the following question.... "Does anyone out there have any real data on the difference in loss between a standard gold plated SMA connector and a stainless steel one at frequencies of 1296MHz and 2304 MHz. I have a built in prejudice against using them at these frequencies but it is only that, I don't have any data."

Chris,G4DGU, made the following sensible comment: "My qualitative view is that there's not a huge penalty - perhaps 0.02 - 0.05dB extra loss at 10GHz. My reasoning is that the centre conductor (which has the highest current density, of course) is gold-plated anyhow, and that losses in the outer are mitigated by the much lower current density. I've not seen any evidence of excessive temperature rise in a pair of mated s/s SMAs carrying 40W at 10.3GHz."

It proved difficult to find one for one comparisons and most SMAs are in fact stainless steel plated with gold. I found two manufacturers websites quoting figures for stainless steel gold plated of (0.03x square root of frequency in GHz) dB... but I couldn't find any figures that were clearly and definitely for non plated stainless steel. For comparison, 10cm of 0.141" coax is (0.038 x square root of the frequency in GHz) dB I found some other interesting data however.

A hermetically sealed connector is (0.05 x square root of the frequency in GHz) dB and brass SMA connectors are quoted as (0.06x square root of the frequency GHz) dB, clearly these are to be avoided. (If you can't tell the difference between brass and gold then try "Alchemist" in your Yellow Pages).

The Radiall website at
www.radiall.com/vdocportal/radiall.jsp
(then follow SMA and Literature, page 11-12)...
has some interesting data . It quotes straight SMA connectors as 0.02dB flat from 1 to 10GHz (which doesn't seem right, it must vary as the square root of frequency) but quotes a right angle connector as 0.04dB at 1GHz, 0.05dB at 2.4 GHz and 0.09dB at 12.4 GHz. I suspect that much of this is a VSWR contribution but it would seem that a minimum length 0.141" cable with two straight connectors is a better solution than a right angle connector. Also in the Radiall data is a useful chart of power rating vs. frequency for SMA connectors.

Now you might think that none of this really matters but at 1296/2320MHz with a low noise dish and feed then, if you have a NF of 0.35dB, changing the loss in front of it from 0.1 to 0.2dB will worsen the receive sensitivity by 0.5dB (see VK3UM eme calc programme at: **web.telia.com/~u92010241download.htm**)

10GHz Losses in Connections between circuits

André Jamet,F9HX and Michel Métroz, F1EER

What's the problem ?

We need to have connections between circuits in the chain of a transmission-reception system, especially for the antenna. Their losses are very penalising as they can ruin the output power as well the actual receive sensitivity. Between intermediate circuits, it is less crucial since the losses can usually be compensated for. As the frequency is increased, interconnection losses are higher due to the increase in component losses. We have to take care to minimise interconnections lengths and to not use lousy components!

What arrangements do we have to do ?

At 10GHz and a fortiori, the rule above is to use wave guides as they have minimal losses as shown in **Table 1** which also gives losses for commonly used conductors and components, as well their SWRs.

The data is valid for **new/in good condition components** but is not generally the case for second hand ones. It's wise to invest in new connectors for the more critical connections in the same way as it is to use high quality capacitors (typically ATC ones) for the PA output and the preamplifier input.

If the connection to the parabolic feed is made by a wave guide and if we are lucky to get a waveguide relay, we will obtain the minimum of losses. If the PA and the preamplifier are also wave guide connected, we will get the best solution but if they are SMA connected we will need a transition to go the antenna wave guide.

If we do not have that kind of relay but only a coaxial one, we'll need SMA connectors to go to the PA and the preamplifier. In order to have a short connection we can use a straight coupler SMA-SMA. If we need a longer connection and angles we'll need a very good quality cable (for example a semi-rigid one) with two connectors. So that we have two routes, one for transmission and one for reception, we can compute the total losses from the individual components losses.

SWR effect

Standing waves are created by a lack of impedance matching between various circuits and by components themselves. When the SWR increases significantly the losses also increase, as shown in **Table 2.**

Moreover, we are not sure that the antenna, the PA and the preamplifier are well optimised to comply with a purely resistive 50 ohms impedance. Very often, the preamplifier requires a reactive source to get the weaker noise factor. That means a 3 to 1.5 SWR (return loss 6 to 14 dB). An advantage of the all wave guide connection is the chance to put independent matching screws in both transmission and reception routes so we can adjust them separately (which is impossible when the wave guide is common for both routes and in the case of coaxial connections). Of course it is essential to use very finely adjustable matching screws, without bad contacts and locked up in the best position.

Measurements

In order to verify data given by the manufacturers, some measurements were done at 10GHz on several components at our disposal. Loss measure-

Table 1

COMPONENT	DATA at 10 GHz
wave guide W90/R100 (copper)	-0.12dB/m SWR ≤1.03
semi-rigid coax .141 RG 402 Ø 3.58mm	-1.52dB/m SWR ≤1.04
SHF 3 cable	-1.88dB/m
RG 58	-1.50dB/m @ 3GHz
RG 174 / KX3 cable	-0.56dB/m @ 400MHz
SMA straight connector semi-rigid cable	-0.10dB SWR 1.15
SMA straight with flexible cable	-0.20dB SWR 1.20
SMA right angle with semi-rigid cable	-0.15dB SWR1.20
SMA right angle with flexible cable	-0.25dB SWR 1.35
relay SPDT with SMA females	≤0.40dB SWR ≤1.40 isolation ≥60 dB
relay SPDT in wave guide	-0.05dB SWR ≤1.10 isolation ≥70 dB
flange for W90 guide	-0.01dB
adapter guide/SMA f right angle	≤-1.10dB SWR ≤1.5
adapter guide/N female	-0.30dB SWR 1.15
straight coupler SMA m/m	-0.10dB SWR 1.05
straight coupler SMA f/f	-0.20dB SWR 1.1
right angle coupler SMA m/f	-0.25dB SWR 1.5
right angle coupler SMA f/f	-0.25dB SWR 1.7
transition wave guide/SMA f straight	-0.30dB SWR 1.15

ment was simply done with a 10GHz source and an 18GHz spectrum analyser (HP141T). The validity of the method was tested by the measurement of a 3 dB attenuator. Others measurements for losses and SWR were done with a vector analyser, ANRITSU 37269A, (40GHz) and some with a test bench equipped with a sweep oscillator (HP8350B) with a 2-18GHz HP 86290A, an 18GHz detector (HP11664) and an 18GHz SWR bridge (Wiltron 87A50-1).

The results are given in **Table 3.** In addition, curves for two SMA attenuators were drawn up to 40GHz. It

Table 2

SWR	losses in dB									
1	0.2	0.5	1	1.5	2	2.5	3	3.5	4	5	6
1.5	0.2	0.5	1	1.5	2.1	2.65	3.15	3.65	4.2	5.2	6.2
2	0.28	0.65	1.2	1.8	2.3	2.9	3.4	3.9	4.4	5.5	6.5
2.5	0.3	0.7	1.3	1.9	2.5	3	3.6	4.2	4.7	5.8	6.8
3	0.32	0.8	1.5	2.1	2.8	3.4	4	4.5	5.1	6.3	7.3
4	0.4	0.9	1.8	2.5	3.3	3.9	4.5	5.1	5.4	6.7	7.8
5	0.5	1.2	2.2	3	3.8	4.5	5	5.5	6.2	7.3	8.4
10	1.2	2	3.5	5	5.5	6.2	7	7.5	8.2	9.5	10.6

is obvious that, above 18GHz, losses and SWR grow rapidly and this is normal for this kind of connector.

EXAMPLE

Two examples of connections will show how we can lose decibels and QSOs !

First case: wave guide entirely

- connection feed, parabolic to SPDT relay: 10cm waveguide WR90 0.01dB flange 0.01dB
- SPDT wave guide relay 0.05dB
- connection to PA or preamplifier: 10 cm wave guide WR 90 0.01dB flange 0.01dB

total losses: **0.09dB**

Second case: SPDT SMA relay

- connection feed parabolic to relay: 10 cm wave guide WR90 0.01 dB
- adapter wave guide/SMA 0.30 dB
- SPDT SMA relay 0.40 dB
- connection to PA or preamplifier: 10 cm coaxial cable .141 0.20 dB

-**total losses: 1.21 dB**

The 1.12 dB difference between the above cases could be regarded as negligible but if we make an effort to afford a 0.7 dB noise factor preamplifier, home made or bought, we will have an 1.98 dB in the second case. During transmission, the difference should be to appear more negligible. However for a 10W PA (+40 dBm) we start from + 39.91 dBm to + 38.99 dBm that is to say 1.87W excess loss. In practice **it is certain that the coaxial connection case is worse than already calculated**. A high SWR will increases losses and the difference with the wave guide solution. The SWR of the SMA relay alone can reach 1.4. Moreover, if the SMA connectors are not tightened with a torque wrench, losses could be worse. Coaxial cables can be damaged if they are bent with too short a radius. So losses can be fatal in relation to a difficult QSO, especially in EME applications.

Conclusion

Microwave watts are very expensive (that it is currently known!) owing to the price of transistors. In reception, lowest noise factors are expensive too. So do not spill precious decibels because they make or break a possible QSO.

It is obvious that those remarks apply to other microwave bands than 10GHz, as the problem grows very fast when the frequency is increased.

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Appendix

The additional losses dues to the SWR is calculated (see ARRL HANDBOOK) by :

total losses for a line = 10 log (B2 – C2) / B (1 – C2)

where :

B = 10 (line loss in dB) / 10
C = (SWR – 1) / (SWR + 1)
SWR is that we can measure at the load, for example :

For a 2 dB line losses when the SWR = 1
if the SWR is reaching 5, we have :

B = 10 (2/10) = 1.585 et C= (5-1)/(5+1) = 0.6666 and :

losses with SWR = 5 : 10 log(1.5852– 0.6662) / (1.585 (1-0.6662)
= 3.74 dB

Table 3

COMPONENT	Measured at 10 GHz
straight coupler SMA m/m # 1	-0.07dB SWR 1.04
straight coupler SMA m/m # 2	-0.08dB SWR 1.1
straight coupler SMA m/f stain steel	-0.07dB SWR 1.02
straight coupler SMA m/f gilded brass	-0.06dB SWR 1.02
right angle coupler SMA f/f	-0.25dB SWR 1.45
Adapter guide/SMA f (homemade with SWR screws not adjusted)	-3.30dB SWR 1.48
As above (but screws adjusted)	-0.5dB
adapter wave guide/SMA f (on a side)	-0.28dB SWR 1.15
adapter wave guide/N f (at the end)	-0.27dB SWR 1.03
black flexible cable KX3/RG 174 325mm SMA m	-5.08dB SWR 1.86
As above but 192 mm -3dB brown flexible cable PTFE SR142B 500 mm Ø 5 mm with two SMA male	-0.85dB SWR 1.06
re-shapable cable .141 red braid 925 mm Suhner SUCOFORM 141 CUPE with right angle SMA m and SMA N f panel mount	-1.52dB SWR 1.12
Re-shapable.141 59,5 mm with SMA m and SMA m	-0.91dB SWR 1.07
conformable cable .085 237 mm with two SMA m	-0.70dB SWR 1.19
semi-rigid .141 6 bends 220 mm with SMA m, f	-0.32dB SWR 1.04
semi-rigid .085 (5 bends) 347 mm with SMA m, f	-0.82dB SWR 1.1
semi-rigid .085 (6 bends) 374 mm with SMA m, f	-0.92dB SWR 1.14
semi-rigid .141 65 mm with two SMA m	-0.50dB SWR 1.18
semi-rigid .141 72 mm with two SMA m	-0.12dB SWR 1.12
semi-rigid .141 273 mm with two SMA m	-0.41dB SWR 1.04
semi-rigid .141 120 mm ? shaped with two SMA m	-0.80dB SWR 1.09
semi-rigid .141 300 mm trombone shaped with right angle SMA m and right angle SMA f	-0.56dB SWR 1.22

A New (?) SMA to microstrip Transition

© Charlie Suckling, G3WDG

A new method of connecting the ground plane of a microstrip to an SMA connector has been developed as part of the effort to design some 2.3GHz power amplifiers.

Normally the body of the SMA connector is soldered to the ground plane of the microstrip directly, as shown in Fig 1(a). This method is used in the majority of microwave pcb designs. However this method causes some problems in power amplifiers when the pcb is to be mounted either flat on a heatsink, or just above a heatsink, in that the solder fillet on the ground plane side gets in the way, and the heatsink needs to be filed to accommodate it. Also, the connector has to be soldered to the board before the board is fitted to the heatsink, which causes major mechanical difficulties if the connector is not soldered perfectly square to the board, or if the heatsink is not exactly the same length as the pcb.

To get round these difficulties, a new method of grounding the body of the connector has been developed, where grounded pads of metal are provided on the top side of the pcb, and these are soldered to the connector with no connection on the ground plane side. In the prototype, three vero pins were used to ground each pad. The configuration is shown in Fig 1(b).

The performance of the new transition is shown. The performance of the conventional method is also given for comparison. In both cases, the microstrip was a 20mm long piece of 50 ohm line etched on 0.79mm thick material.

The launch to the other end of the line was via a conventionally mounted SMA connector. It can be seen that the new transition has almost the same performance as the conventional type, having better than 30dB return loss up to about 2.8GHz and better than 20db return loss to nearly 7GHz.

This type of transition is being incorporated into the latest prototypes of my 1W and 10W 2.3GHz power amplifiers.

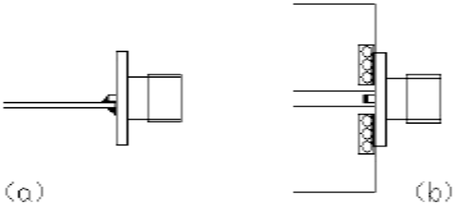
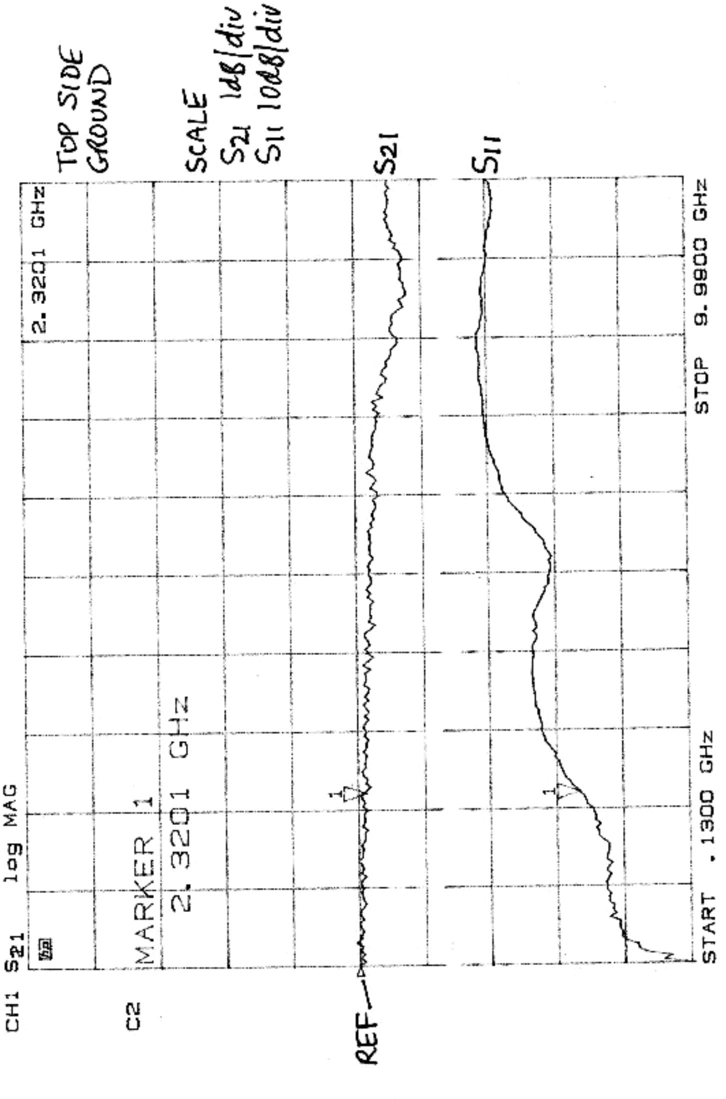


Figure 1


CH1 S21 log MAG
C2
MARKER 1
2.3201 GHz
2.3201 GHz
1
1
START .1300 GHz
STOP 9.9800 GHz
REF
S21
S11
TOP SIDE
GROUND
SCALE
S21 1dB/div
S11 10dB/div

Tips for Operating a Multiband Station

Uffe Lindhardt, PA5DD

Introduction

Being QRV on microwaves can involve being QRV at 5,6 or 7 different bands. Contests and band openings often requires fast QSY between these bands, as well as simultaneous operation on more than one band at a time (e.g. talkback). To fulfil these requirements puts high demands on the station design. This article tries to give some hints as to how this can be done.

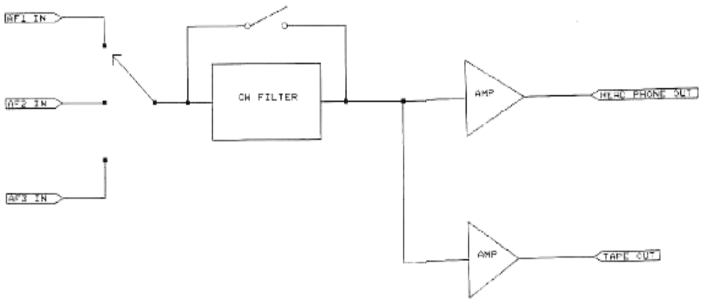
Audio and CW keying

In order to be simultaneously QRV on different bands, more than one transceiver is needed. As the choice of talkback varies (144 MHz, 432 MHz, 1296 MHz etc.) even more transceivers could be needed.

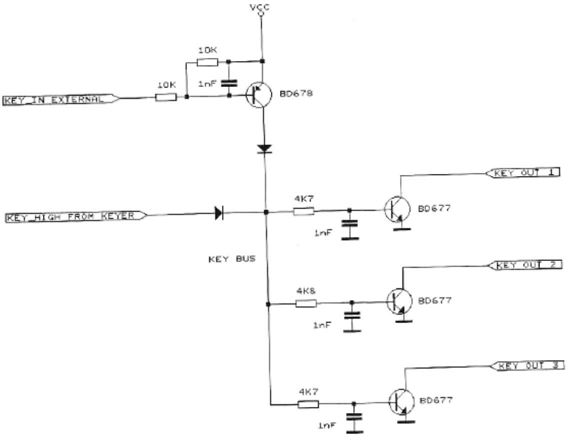
This naturally gives some switching problems when it comes to one headphone and one CW keyer.

For the audio part I have made a “Headphone amplifier box” which takes the standard 200mV AF output available on most commercial transceivers as input. For the sake of standardisation I have equipped all my transceivers (6 in all covering 1.8 MHz trough 24 GHz) with phono plugs for the AF output. Inside the box I have a switch to select one on the AF inputs. The input is then amplified before being fed to the headphone. Result: I do not have to plug my headphones in and out all the time. Taking advantage of the concentration of audio signals in the amplifier box, I have fitted a switchable audio CW filter and an isolated tape recorder / sound card output in the AF line going into the amplifier.

For CW keying I have chosen to key all transceivers in parallel. I am assuming, that only one transceiver will be transmitting at a time, so do not use break-in, or all your transceivers will be transmitting simultaneously. On the other hand this system removes any switching requirement for the CW key. In order to achieve keying in parallel and having isolation between the keying inputs of the transceivers, I have chosen to make a keying serial "bus", which is active high. This bus keys a series of NPN switching transistors, one for each transceiver. The bus is driven from my electronic keyer, which has an active high (inverse) output. Extra inputs to the bus for a hand key or PC interface can be added by means of an inverting circuit (e.g. PNP transistor).



This is the circuit diagram:



IF Switching

In order to bring down the transceiver count, it is natural to use the same transceiver for more microwave bands I prefer to use separate RX & TX cables for the IF connection to my microwave transverters. This "dual line" approach makes it easier to introduce separate attenuation in the RX or in the TX branch, when there is a need to adjust the signal levels. It also gives more transparency during fault detection on a mast mounted transverter. Furthermore I use split RX & TX cables also on the lower bands, so my transceivers already have separate RX & TX connectors.

I order to speed up the process of QSYing from one band to another, I have built some IF switches. Since separate switches are need in each branch (RX & TX), and maybe a third switch for the TX control signal, I have chosen low cost DC relays. These relays do not provide too much isolation on 144/432 MHz, but that is of less importance. I am using small "sugar cube" relays, which have a metal cover, that can be used for grounding. These relays can be found surplus at reasonable prices, and by using more relays a split of 2,4 or more IF outputs can be provided. If you are using a pre-amp and a PA on the fundamental band of the transceiver, then even that band could be connected to one of the IF outputs.

The relays could be internal or external to the transceiver, but it is an advantage, if the switching can be controlled from the transceiver. In my ICOM 402 I use the S-meter light switch voltage to switch the relays as well, this gives a good visual indication of which band I am on (e.g. lights off is 6cm, and lights on is 3cm).

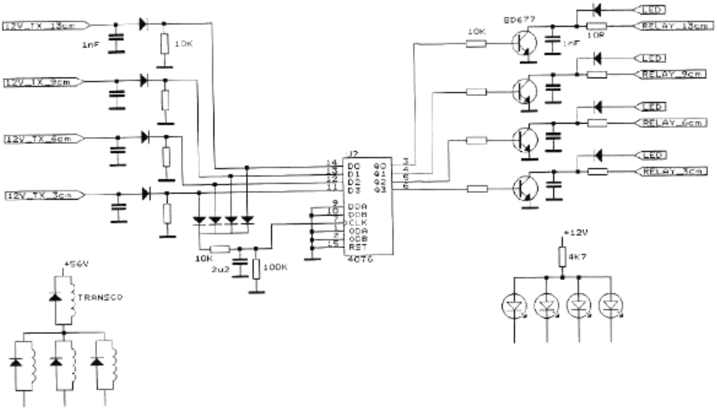
Multiband Antennas

In order to make an efficient use of antenna mast space, it becomes desirable to make use of multiband antennas, e.g. a parabolic dish with a multi band feed. If at the same time the transverters have been placed in the mast to minimize cable losses, a need for switching between the antenna and the transverters occurs. Fortunately multi pole microwave relays are available e.g. TRANSCO

14300, which is a SP4T relay.
In order to make the switching as seamless to the operator as possible, I decided to make some control logic, that switches the relay to a given band when the transverter of that band is switched to TX.

Afterwards the relay stays in that position until another transverter is switched to TX. With this system the operator simply has to flick the PTT on a given band to acquire a connection to the multi band antenna.

The control logic was realised with D-flip/flops, connecting the TX/RX switching signals (e.g. 12V_TX) output signals of the transceivers to the inputs, and using the RX-to-TX transition of any of the signals to clock the flip/flops. The schematic below shows a system for four bands using the TRANSCO relay.



A Tracking Modulator/Detector System for Measuring Microwave Cavities

John A. Share, G3OKA

A simple method of determining the resonant frequency of a newly made cavity is to excite it with a low level signal from a generator and to look for a peak in the detector output. A typical microwave signal generator will deliver +10dbm and when lightly coupled to the test cavity the resultant signal from the detector is probably less than 1 mV. Amplifying this direct current signal can pose numerous problems, an established solution is to amplitude modulate the signal generator and measure the amplitude of the modulation from the detector. This is an ac signal and being at audio frequency its amplification poses few problems. A fundamental problem does exist in that the detector output contains not only the required audio frequency but also noise and 50Hz. Some filtering is therefore essential.

The ready availability of integrated Active Filter devices, and the ease with which they can be configured for centre frequency and selectivity make them an obvious choice. Their centre frequency is controlled by their clocking rate, the devices used had an on chip oscillator that required only two external components, or they could be clocked from an external source. This clock was also used to generate the modulation, if it varied then the detector filter tracked the modulation. High selectivity in the detector filter could then be used to improve the signal to noise ratio. A number of MF8CCN and 741 Operational Amplifiers were immediately available and a circuit was designed using these devices (**figure1**). There are other, probably better, devices that could be used. All the circuitry is at audio frequency and there are few constraints on the amplifier devices, circuit layout or construction.

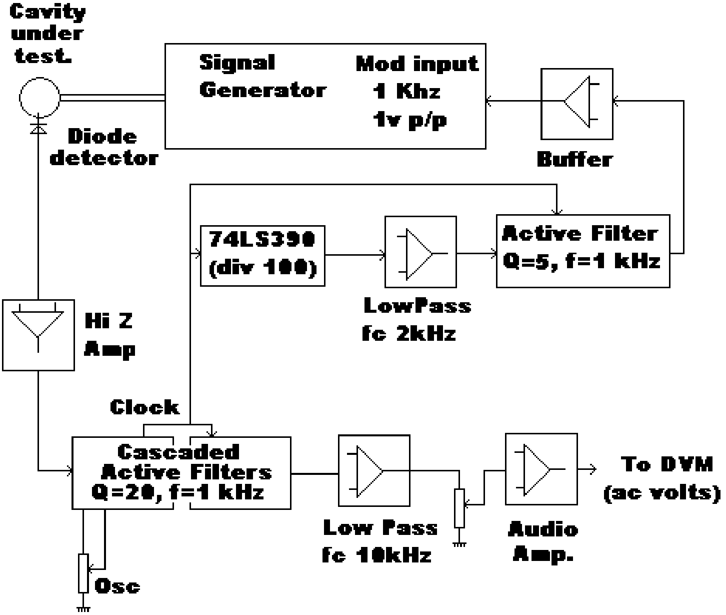
The cascaded detector Active Filters have a hardwired Q of 20, this was found to be more than adequate, and the centre frequency was selected for a nominal 1 kHz, clock rate is at one hundred times the centre frequency. This clock is also divided in a dual decade divider (73LS390) generating a 1 kHz square wave. A simple low pass filter with a cut off of 2kHz removes the odd order harmonics from the divider output and passes a recognisable sine wave into the modulation Active Filter.

This is hardwired for a relatively low Q and its output is a sine wave of good purity. The modulation frequency and the detector filter centre frequency are thus locked.

The MF8CCN has a relatively low input impedance, as such it excessively loaded the detector and a high input impedance amplifier was laced between the detector and the filter. The design constraints on this stage are negligible, it needs to have an input impedance of >1 Megohm and a bandwidth of a few kHz and was built around three 741 Op amps with a nominal gain of ten.

The Low Pass filter following the Detector Filter is necessary to remove switching transients at clock frequency that are superimposed on the output signal. It is possible that a passive R/C filter would remove these spikes but an active stage eliminates them completely. The Audio Amplifier simply increases the amplitude to a convenient level for the Digital Volt Meter. The Keithley 199 DMM that was used as the output indicator is able to measure ac signals at Audio and there was no need to rectify the signal.

This circuitry has been used extensively in the measurement of cylindrical cavities developed for use at 9 GHz and 14 GHz.



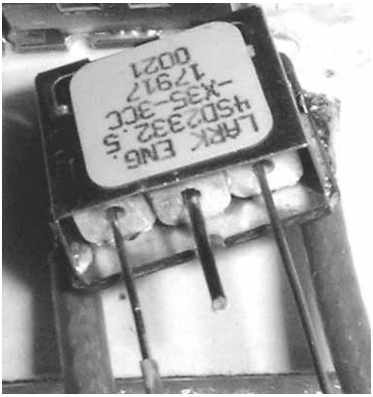
Tweaking Commercial Ceramic Filters

Kent Britain WA5VJB

WB0TEM picked up a large number of these 2332 MHz Ceramic Filters. He then pulled a dirty trick by giving me 50 of them!

They will work, as-is, at 2304MHz, but the response is falling fast and then you need to try a few to find a good one.

Marc was hoping I could figure out a way to re-tweak them. It seems that you can!



Find some solid copper wire that just fits back down the centre opening of the ceramic element, something in the #20-#24 range. The same size wire might not fit all the elements.

Now tweak away. If you go all the way in, the element will short out, but there is a space near the bottom where you're retuning the filter element. I got these to go down to 2275 MHz, a drop of over 50MHz with little trouble. Next, a dab of RTV (or my favourite, Liquid Nails) to hold the wires in place. Then trim off the excess wire after the glue dries.

Super Glue wicked all though the filter making a mess …. I don't recommend Super Glues!

Static Sensitive Devices and Fingers
Static Discharge Experiences

Julian G3TFR

I am all too painfully aware of the effects of major static discharge, which can be encountered in some homes, or in office areas, even of electronics factories. One office carpet I knew would, on a dry day, raise a body to some tens of kV after only a few steps across it, resulting in a powerful jolt and involuntary, minor, muscle spasm, on touching the first filing cabinet or earthed metalwork. I tried a couple of measures to protect myself from this, carrying either a 10Mohm resistor to damp the oscillatory discharge, or a coin or key to spread it over a greater skin area to reduce the current density. A good 'crack' would then be obtained, with little finger-pain, though the arm muscles would still spasm.

It intrigued me that some people appeared quite insensible to such discharges - I would notice from the corner of the eye that someone was approaching and about to lean against my filing cabinet, and while I was still trying to formulate a verbal warning, they would put out a hand to touch the cabinet, just as they began to speak: "I say, could you (KRRACK!!!) find me - er, what was that funny noise? - er, find me that purchase specification..." They literally hadn't felt a thing, just heard it, whereas I would have twitched and jumped, and would kick the filing cabinet for its audacity. So, I was more keenly aware than some, of the unstoppable discharge current of such a charge, and obviously no semiconductor device could survive such a discharge. The charge could easily be carried from the office to the bench. Merely pulling off an acrylic pullover while at the bench could prejudice everything in sight. Latterly, my procedure for protection from pullover-static is to touch, with the palms of the hands, a wall, preferably plastered brickwork, even if vinyl-emulsioned, which will safely discharge one in a second, without sensation.

PROBLEMS IN THE KITCHEN

'Laminate' wood-effect flooring also causes static charge accumulation, and a friend has frequent trouble from this in her kitchen. I noticed her FM radio on the worktop seemed to have become very insensitive, so I undertook to check the first RF transistor. I thought "if only they had bothered to isolate the device from the telescopic rod with a small-value series capacitor, or a small-value shunt RFC" and was then surprised to find both were fitted, but could not protect the bipolar device from the dreaded static.

ANTI-STATIC PRACTICE IN INDUSTRY

Most (all?) industrial electronics assembly facilities nowadays enforce strict static protection measures, including: anti-static clothing, tools, packaging, storage, stationery, conductive flooring and conductive over-shoes, the flooring often consisting of a vinyl surface, but with flecks of black conductive elastomer connecting through to a conductive/resistive backing earth leads for personnel, connecting either to wrist straps or to studs on the conductive clothing which in turn have conductive elastic sleeves static-dissipative (capacitive) or conductive worktops, connected to ground

Note that all conductive paths such as ground straps, worktops and floors, are protected by high-value resistors or by highly resistive materials. A strict regime of tests and safety checks is employed and documented, including daily checks on all wrist straps. No operatives are allowed to work on the benches without these measures, and entry to some factory zones may be restricted.

MISATTRIBUTION

I am always contemptuous of attempts to make engineers wear wrist straps under all circumstances when working at the bench. They of all people should be aware of whether or not that is necessary, just as the competent driver should be aware whether or not it is always necessary to use the indicators. I would say that even rudimentary measures will eliminate the risk of static damage, and though I have often seen device failures at work which are attributed to 'static damage' I believe that such attributions are usually superstitious. Closer inspection will almost invariably reveal a 'true' cause. Microwave power FETs may destructively 'hoot' while being shimmed, sensitive Schottky diodes may be blown by induced earth-loop currents, internal bonds in MMICs may fail during soldering, or even spontaneously if manufacturing standards are poor.

A WARNING

I was recently asked to work on some microwave Doppler units (long-established microwavers will recall these things with affection - see G3PHO's webpage article, "A History of 10GHz"). They consist principally of a bit of waveguide containing a Gunn diode which will always measure dead short on a DVM, but isn't, and one or two mixer diodes, which are seriously delicate, with reverse breakdown ratings around two to three volts - you can destroy them with the wrong DVM. I was seeing a large percentage of dead diodes, but found one good one and measured its sensitivity and ten minutes later, to my considerable chagrin, found it to be dead. I had even been using the blasted wrist strap, to please the client. Now, preparatory to starting this work, I had spent a day or more rebuilding the pair of fine old lab PSUs I intended to use, paying particular attention to replacing the mains leads, and observing 'best practice' by connecting the earth conductor using a crimp terminal, with star washer and lock-nut. They had MK plug-tops, which are the only ones which do not work loose their connections. Checking after the disaster, my DVM showed-up 100V AC between the two PSU outputs, though the static sensitive fingers had not felt anything. This turned out to be due to slack screw terminals in the mains socket boards under the bench. Microwave constructors, go and ponder your own vulnerabilities!

MY RECOMMENDATIONS

1) Run a safety earth wire between every individual piece of equipment used on the relevant bench, connected to earth terminals where provided on the equipment, or to any convenient case screw.
2) When making periodic checks on earth connections, do not use a DVM as the response time is too slow to catch short intermittent breaks. Use a fast-responding buzzer, or a lamp-bulb and battery and waggle the cables.
3) Assume that the earth continuity of all mains sockets is suspect until checked and, once checked, assume it will fail sometime in the future.
4) Appreciate that copper wire will slowly deform under high pressure, which is why the cable connections on sockets, and non-MK plugs, nearly always loosen with time.

Soldering ‘pipe cap’ filters on a DEM 10GHz Transverter board

Bob, WA5YWC

I recently purchased the DEM (DownEast Microwave) 10GHz transverter through the North Texas Microwave Society project.

During the assembly of the transverter, I came to the part about soldering the pipe cap filters to the pcb with a propane torch. I was a little uncomfortable with this method without first experimenting with a sample pipe cap.

First, I cleaned a copper cap with Scotchbrite and decided to tin the open end with solder. The results, in my estimation, were unacceptable. The pipe cap didn't tin very well due to excessive heat from the torch and was all black and nasty. I was not about to destroy my transverter board with poor soldering and excessive heat from a torch!

I came up with a much friendlier solution: use a heat gun. I purchased a Paladin heat gun at the local electronics store. It comes with three different removable nozzles and variable heat range. I also have a cheap, metal soup ladle (found in any grocery store kitchen gadget isle) I use for ladling hot solder for making fishing sinkers. By unwinding about six or seven feet of solder from the roll and placing it in the ladle, I heat the solder with the heat gun. It melts very quickly.

Next, I heat a shiny pipe cap with the heat gun and dip the open end in solder paste. Quickly, I put the open end in the hot solder and apply more hot air from the heat gun. The solder tins the end of the pipe cap very uniformly, inside and out, to a height of approximately one-eighth inch. I clean the pipe cap with alcohol and it is still bright.

With this method, I prepared all nine pipe caps for the transverter. Now, it was time to solder them to the board.

I followed DEM's instructions to mount the board to the aluminium pallet. With the heat gun, I pre-heated the pallet and board assembly. Taking a prepared pipe cap, I heated it for a few second to get the solder flowing. Then, I dipped the cap into the solder paste and dropped it in the appropriate hole in the pallet. Now, applying heat in a circular motion, and applying a small amount of solder in the corners of the pallet holes, the pipe cap was uniformly soldered in place. I continued with this method until all pipe caps were in place.

After the assembly cooled, I removed the board from the pallet and found nine, perfectly soldered pipe caps.

I cleaned the assembly in alcohol and baked it in the sun for a few hours.

I hope this information will be helpful for those intimidated by using a propane torch on this great microwave project.

Bondage …

© Lehane Kellett, G8KMH

Yes, you are reading BackScatter and you haven't picked up the wrong book! I guess I should really have titled it Bonding but that may have been as bad.

So let's straighten things out! I'm talking here about wire bonding, the 21st Century soldering iron. I was prompted to write this as I prepared to start some commercial prototyping and also many will have seen DB6NT's 47GHz amplifier at Martlesham.

We are all now familiar with the use of SMD components and these, at least in commercial usage, get ever smaller. 0402 is a pretty common passive device size in many portable devices and SOT363 for actives. How do you get smaller? And why? One of the reasons is to get even more into one area and the second is that at even higher frequencies (>c.20GHz) lead lengths and package parasitics kill the device performance. So the only thing to do is to connect directly from the active device to the circuit lines/microstrip. And for this you need to bond to the die and the track. Sounds easy?! Hardly, typical microwave devices (MMICs) are 1.6mm x 0.8mm and individual FETs and capacitors even smaller. So as you can imagine the wire used is extremely fine – usually 25um (1mil) and made of exceptionally pure gold, although aluminium is used in other branches of microelectronics. As an aside, it is also exceptionally expensive – 20m of 1mil on a spool is $375. So we have the wire, lets solder it to the chip….

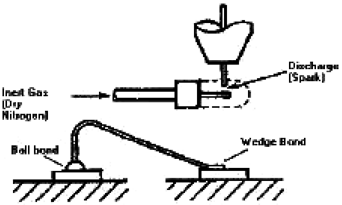
You first! My Acme soldering iron tip is a tad large. It just isn't feasible to use soldering techniques (OK, the smartass who said Indium solder gets a Gold Star) or even use conductive epoxy. The solution, which has been around for a very long time, is to weld the wire to the pads on the die and to the tracks. It really is welding too, the wire is pushed against the pad/track and enough energy applied to get the two to alloy and not enough that they melt completely – or destroy the die. Usually the materials are identical too which is why you see so much Gold inside commercial equipment.

So the trick is to have a machine that can spool the wire, hold it down and apply energy, then do it again and finally break the wire off. It may sound like a bit of a tall order but they exist in many shapes and sizes (and costs!) and are known as wire bonding machines, no surprise really. Well known names include Hughes/Westbond, Kulicke & Soffa and Hybond (the picture is a Hybond 572A at Queens University, Belfast)

The trickiest bits are getting the force right and applying the energy. The first is solved with some simple mechanics and in modern machines, strain gauges. And with the second there are two main methods of bonding – thermosonic/ball bonding and thermocompression bonding.

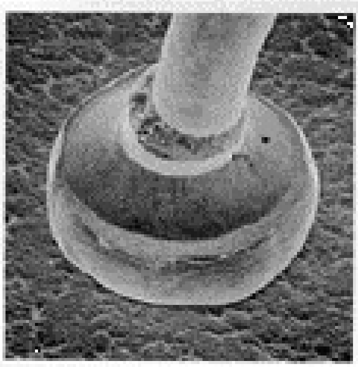
Ball bonding relies upon creating a blob of molten metal just a fraction of a second before the wire makes contact with the preheated die/track/package frame and then it is pressed down. Why pre-heated? If you think about it, putting a hot blob against a cold piece of metal isn't going to give an alloyed joint. This method of bonding is ideal for some devices and probably a large proportion of silicon transistors and IC's use this technique. The magic TLA here is EFO – Electronic Flame Off. The reader is left to other interpretations.

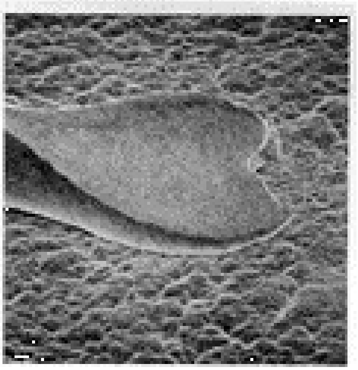
After the first bond is made to the die then the wire is pulled through the bonding tool and then pressed onto the track/frame and ultrasonic energy used to make the weld.



Ball bonding isn't used much with microwave devices. Thermocompression bonding is the workhorse of microwave electronics and is used for everything from mounting MMICs to linking modules. In this case the tool used is more wedge shaped and the wire (or ribbon) is fed through the tool at an angle—45° or 30° are standard. The wire is then compressed against the pad and a burst of ultrasonic energy used to create the bond. The tool then lifts and wire is fed through the tool until the second bond position is reached and the process is repeated. This may be the final bond or there may be another – this is used when the connection from the MMIC first goes to a decoupling capacitor (Di-Cap) before going to a supply track. The wire is simply broken by clamping the wire as the tool is moved upwards, it then breaks at the weakest point. Heat is provided to the die/assembly on unit known as a work stage. Typically this is heated to 125C. Some processes call for the tool to be also heated in place of, or in combination with, the ultrasonic burst. Every MMIC manufacturer seems to have slightly different bonding parameters. As you may imagine, all of this requires everything to be scrupulously clean, as any residues or particles will cause bond failure. Much commercial processing takes place inside clean rooms but it is possible to maintain sufficient cleanliness inside a filtered air cupboard, usually know as a Laminar Flow Cupboard – many universities do this.

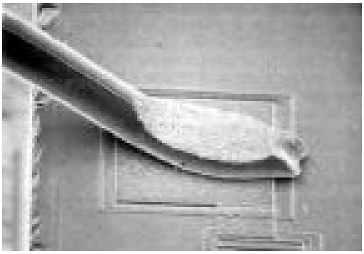
And in order to keep everything clean, the wire, devices, etc. are kept in a Dry Nitrogen atmosphere.





So, you fancy having a go? Well you can certainly pick up good used bonding machines from £800 up but you'll need to add the Nitrogen storage facilities, bonding tools, gold wire and clean room facilities. Say £3-5000 to start.

Makes that soldering iron look like a bargain, doesn't it!



Running Beacons from an Unstable 12 Volt Supply

Andy Talbot, G4JNH

From the outset we aimed to run the GB3SC# family of beacons from a 12 Volt battery backed up supply in order to simplify individual units and make them less susceptible to short mains outages (like using a kettle on site which drops the mains voltage to 120V!). The 2.3 to 5.7GHz beacons were made up as normal, designed to accept the straight 12V input from the battery which was often used directly for DDK004 type multipliers and crystal heaters - we never thought about the implications of interactions between equipments.

The first version of GB3SCX had been built using a mains power supply giving 14V and had proved adequately stable. When converting to 12V operation at its rebuild stage to uprate the PA last year, I noticed that the RF line up was quite susceptible to frequency shifts with supply voltage.

So, from the outset, I had to ensure it was properly regulated and able to cope with whatever a battery supply could throw at it. The DDK004 RF multiplier needed at least 11V to give enough RF output so, assuming a battery voltage that could shift over the range 11 to 15V, it was obvious that simple linear regulators were out. The head unit with its 2.5 to 10GHz multiplier and PA, included its own regulation, so could be left to fend for itself using raw battery supply. So now it was only necessary to supply the DDK004 source. The Farnell / RS catalogues list a plethora of small low cost encapsulated DC / DC converter modules rated from 1 Watt upward. The cost of these increase considerably with power rating so the lower rating we can get away with, the better.

Since a target voltage of 13.8V seemed to be about right, and I wanted to include a linear regulator anyway, a raw converted output of at least 15V is called for. To minimise the rating needed for the DC/DC unit, the isolated output from this can be added to the source voltage, rather than supplying all of it, so the DC/DC module only has to supply a portion of the power needed for the final unit. I already had a 1 Watt rated 12V to 12V converter (NME1212S, Farnell Part No. 178-197, the cheapest they do) from an earlier project and used this in a circuit similar to Figure 1. It could only give 80mA output which rather marginal, but just about sufficient for the DDK004 alone. An LM317* was used as a linear regulator to allow 14V output (7812 regulators jacked up with a couple of diodes are disgusting and should be banned by law!).

With DC/DC converters, EMC is always an issue and I fully expected to have to include plenty of decoupling to remove spikes and noise at the switching frequency. Filtering was built in from the start and proved to be necessary.

Even with the one input and two output stages of filtering as shown, sidebands at -60dB are detectable on the RF output at a few hundred kHz away. All worked as expected and this PSU is the version in current use on GB3SCX. It works properly over the input voltage range 10.5 to 15 Volts (although with slightly reduced power output at the lowest input voltage, down to the design of the 2W PA stage)

As a number of listeners know and complain about, GB3SCC, the 5.76GHz beacon, operates on a time cycle which was initially 40 seconds on and

40 seconds off (now 40 and 20 seconds respectively for a trial period). I noticed that the frequency of GB3SCF on 3.4GHz changed audibly—in fact by about 30 - 40Hz when GB3SCC switched on and off. As this uses raw 12V for some of its stages, the frequency shift is to be expected. What I hadn't bargained for was a slight (about 10Hz) residual shift on GB3SCX in spite of the, now, highly stable supply. Also it wasn't a sharp shift such as that observed on 'SCF. Instead the frequency would change slowly over a few seconds and even appeared to overshoot slightly at times. It looked as if the crystal heater was to blame, changing the frequency as it warmed and cooled with the approximately 150mV supply voltage variation. I haven't measured this effect any more accurately, and the next version of the beacon will use a phase locked approach, not calling for a crystal heater.

I will also need to adopt an uprated DC/DC converter to supply the PLL unit as well as the multiplier. It is not necessary to generate 12V in order to add to a worst case battery input to get 14V, and a lower output is desirable as it allows greater output current for a given rating. In the NML 2 Watt range, output voltages of 5V and 9V are listed. 5V will do for most applications when added to a battery supply of 12V but with a worst case of 10V input (especially if an 'idiot diode' is included in circuit) is too marginal for a guaranteed 14V regulated output. So the 9V NML1209S (Farnell part No. 305-8311) will be used for the next version allowing 220mA of current capability. GB3SCK, on 24GHz, was initially built as a mains powered unit but, after pressure was applied to him, Chris G8BKE converted to 12 Volt operation. To do this he had to incorporate an inverter to supply the 20 Volts needed for its 'brick' PLL oscillator. The first version used a low power packaged DC/DC converter for this task alone, with all the other modules using 'raw' 12V and sometimes regulating this down. Occasionally 24GHz would go completely unstable with the frequency shifting wildly and this was put down to the 5.7GHz unit changing the supply. This time, Chris just took the quick approach just went for a big DC/DC converter that was to hand, generating 20V and regulated down to 12V.

Now the whole beacon complex operates off a battery supply, although the two lowest frequency beacons still have slight frequency shifts as GB3SCC cycles - these may be updated eventually as part of the project to make them all phase locked.

Note for users of LM317 Regulators ...

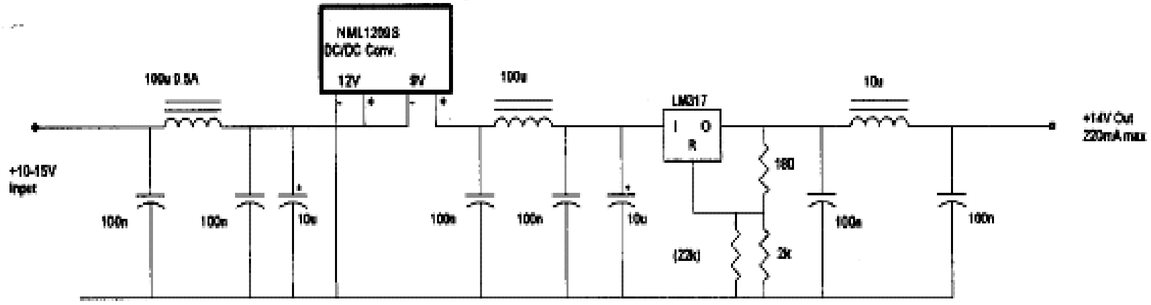
There is an occasional tendency when using the LM317 for users to add decoupling capacitors to the adjust, or feedback, connection as well as between both input and output to ground. Perhaps they feel that as there is no capacitor directly across device pins, the decoupling won't work properly.

This is wrong, even though the device will appear to regulate voltage correctly.

The voltage adjust pin has to carry the high speed feedback signal corresponding to output load changes. Any capacitor added here destroys high speed transient response of the regulation and allows hum and noise to go straight through.

I once saw this error on an LM217 used in the voltage regulation circuitry of a piece of hardware destined to go on a communications satellite. The mistake had gone right through initial bread boarding, and was only detected when the unit dismally failed its EMC tests for conducted susceptibility on the evaluation model.

Figure 1: circuit diagram of the 220mA rated 14V supply to be used for the next generation phase locked GB3SCX source.



Sulphation of Lead Acid Batteries

Steve Davies, G4KNZ

I was wondering if anyone had come across this method (below) for removing lead sulphate crystals? Has anyone actually tried it, or got one of the "conditioners" from VDC? Or does it not really work?

Read on ….

When a battery is improperly charged (over/ under) or allowed to self-discharge, as occurs during storage/ non-use, crystals of lead sulphate (PbS04) build up on the battery's storage plates, preventing the battery from ever being fully charged and therefore able to deliver their full power/ capacity.

Every element has a magnetic moment at a resonant frequency i.e. a point at which the chemical bonds that hold the molecules together to form a crystal can be broken. The resonant frequency of lead sulphate crystals is 3.26MHz. By creating a waveform with the required 3.26MHz frequency, coupled with a very fast rise time and a high amplitude pulse, the sulphation can be broken down.

This approach is used in a battery conditioner from VDC Electronics.

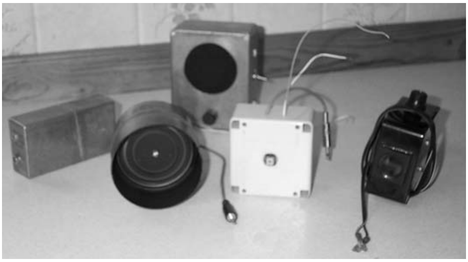
Note: Not every battery is a candidate for reconditioning - for example, failure may be due to mechanical damage caused by vibration or contamination, which has created is "shorted" cells. But If a 12 volt battery has a resting voltage of at least 10.5V and none of the 6 cells are shorted, desulphation of its plates can be accomplished.

High Performance, Low Cost Lightwave RX Front End

Gordon Fiander, G0EWN

The following describes a low cost-high performance lightwave RX that combines parts of the K3PGP design with G0MRF stages.

My first RX was based on G0MRF design but used OPT101 front end. These are still available, via Farnell for around £4.00. The OPT 301 is no longer available as far as I can gather, which is a pity as they have slightly better characteristics than the 101. In practice, however, performance of the 101 and 301 are very similar; a test over a 10km path with Barry, G8AGN TX, and Peter G3PHO (using the OPT301) and myself RX, G0EWN, (using the OPT 101) showed little, if any, difference in performance (both received G8AGN at 59+).



Following the above test earlier this year, I decided to experiment with a PIN diode front end. This was based on the K3PGP design. I found Maplin sold a 5mm package PIN photodiode, the Siemens SFH2030, costing only 79p. I used a 2N3819 in place of the MPF102 and omitted the second stage of the K3PGP design. The output from the 2N3819 was filtered and amplified by G0MRF's audio stages, which use low noise 5534 op amps. TL071/81's are very poor in comparison.

Output from David's unit then goes to an LM386 audio amp set at maximum gain. I compared the Pin diode front end with the OPT101 using a G8AGN weak signal source and found the PIN diode arrangement was much more sensitive than the OPT 101. Tests were made using exactly the same following stages and just substituting front ends. The PIN diode was capable of copying signals the OPT101 could not detect! These tests were made without any optics ahead of the two sensors.

I followed K3PGP's advice of using simple point-to-point wiring and shielding (see photos). By combining the first stage of the K3PGP design and G0MRF's audio filter/amp stages you get a high performance/low cost system that's very easy to construct and is an excellent starting point for dabbling in lightwaves.

G0EWN's lightwave receiver

Backscatter 1st Edition Errata— to be inserted after p347

Figure 1. Circuit Diagram

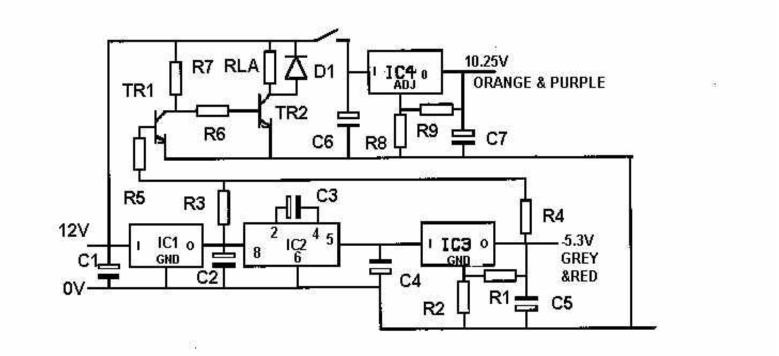


Figure 2. PCB Track Layout

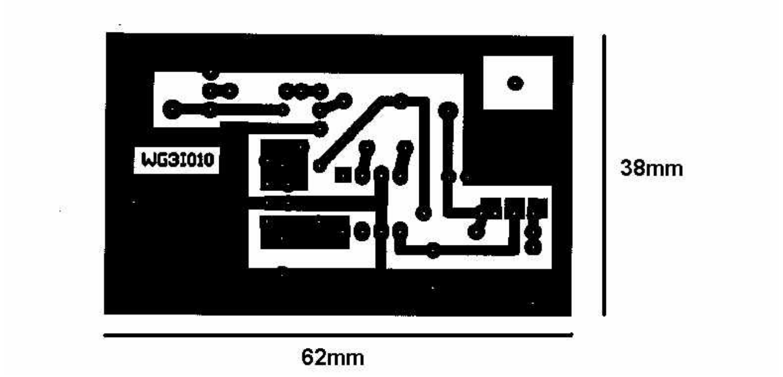


Figure 3. Component Overlay

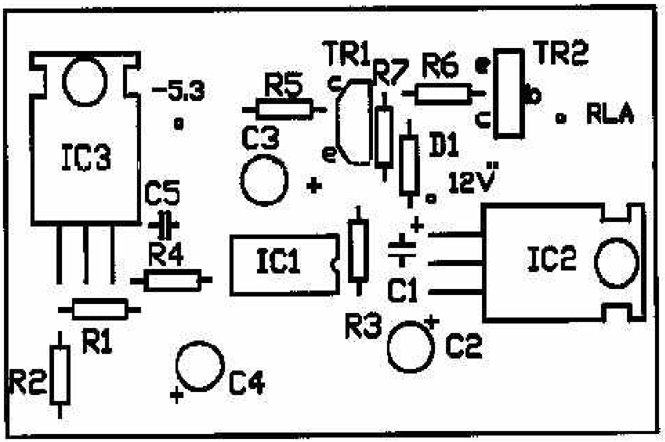


Table 1 Component listing

Component	Value
C1	0.1uF 25V TANT
C2	1uF 16V TANT
C3,C4	22uF 10V TANT
C5	0.1uF Ceramic
C6	10uF 16V TANT
C7	22uF 16V TANT
D1	1N4001
IC1	7808
IC2	LTC1044
IC3	LM337T
IC4	LT1084CT
R1,R6	470
R2,R5,R7	1k2
R3	4k7
R4	3k3
R8	867(820+47 in series)
R9	120
TR1	2N3904 or similar
TR2	BD132 or similar
RL1	DPDT 12V coil 5A contacts